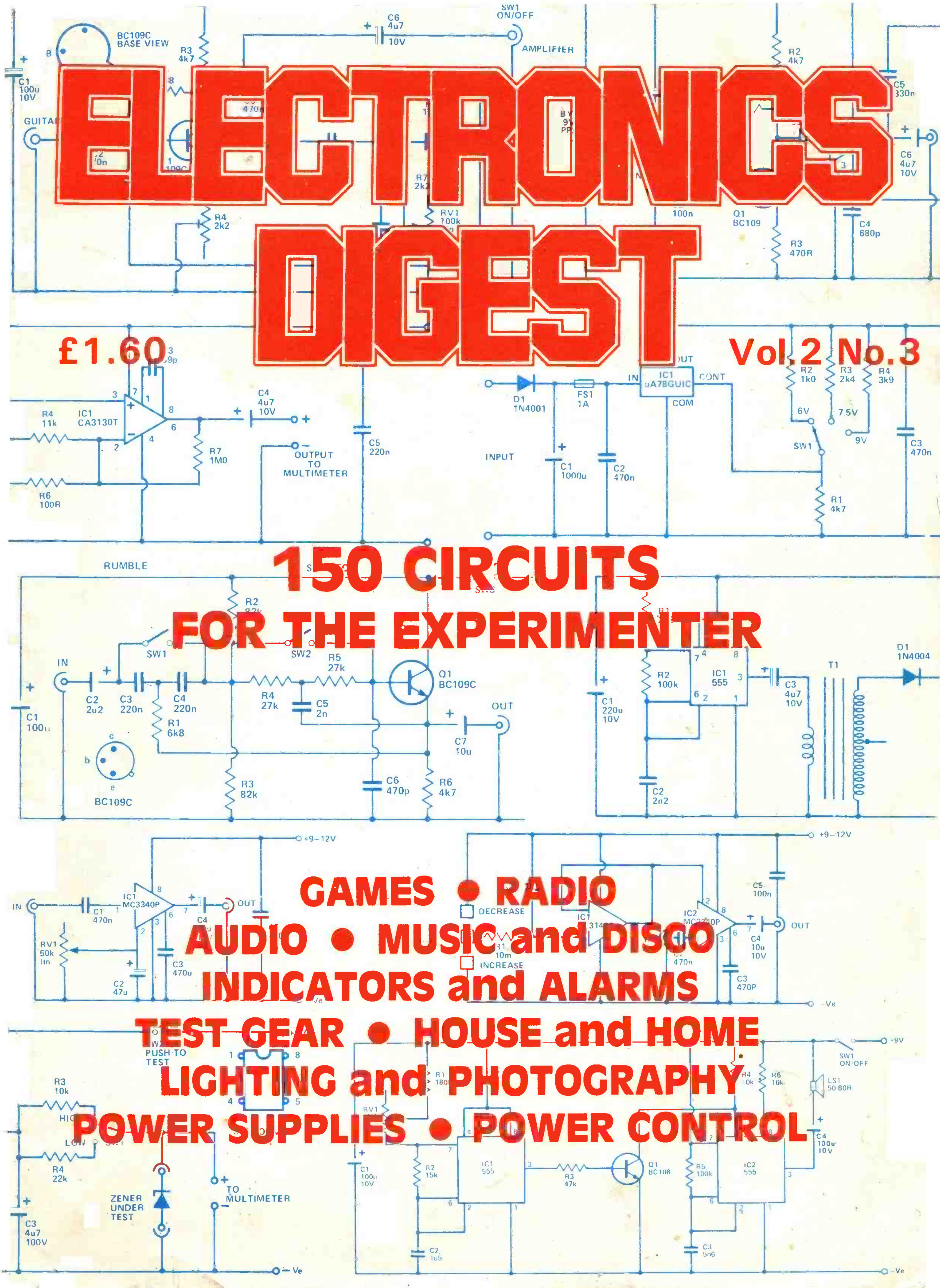
ELECTRONICS DIGEST

£1.60

Vol. 2 No. 3

150 CIRCUITS FOR THE EXPERIMENTER

GAMES • RADIO
AUDIO • MUSIC and DISCO
INDICATORS and ALARMS
TEST GEAR • HOUSE and HOME
LIGHTING and PHOTOGRAPHY
POWER SUPPLIES • POWER CONTROL



OP AMP ABRIDGED PERFORMANCE SPECIFICATION

S = Single D = Dual Q = Quad

Op amp type	Input offset voltage mV	Input bias current nA	Type of input structure	Band-width MHz	Slew rate V/NS	Voltage gain gain dB	Maximum supply voltage V	CMRR dB	Qty	Comments
709	2	300	NPN	1	0.25	90	±18	90	S	***Needs frequency compensation***
307	2	70	NPN	1	0.25	100	±18	90	S	***Internal frequency compensation***
301	2	70	NPN	10	0.5	100	±18	90	S	***Needs frequency compensation***
741	2	80	NPN	1	0.5	106	±18	90	S	***Internal frequency compensation***
748	1	120	NPN	10	0.5	103	±22	90	S	***A decompensated 741***
308	2	1.5	NPN	3	0.5	110	±18	100	S	***Low supply current drain 0.3mA Needs frequency compensation Very low differential input voltage range***
318	4	150	NPN	15	50	106	±20	100	S	***Very low differential input voltage range. Sometimes needs frequency compensation***
747	2	80	NPN	1	0.5	106	±18	90	D	***Internal frequency compensation***
1458	1	80	NPN	1	0.8	103	±18	90	D	***Internal frequency compensation***
4136	0.5	40	PNP	3	1.0	110	±18	100	D	***Low noise***
3900 3401	Current inputs	30	Current sinks	2.5	0.5 20	70	±18	—	Q	***Current balancing amplifier***
324	2	45	PNP	1	0.5	100	+30	70	Q	***Ground sensing inputs Output voltage can go to ground Low power. 0.8mA drain per IC***
3403	2	150	PNP	1	1.2	100	+36	90	Q	***Ground sensing inputs Class AB output Output voltage can go to ground Low power 3mA drain per IC***
348	1	30	NPN	1	0.5	103	±18	90	Q	***Low power 2.4mA drain per IC Class AB output***
RC4739	2	40	PNP	3	1	110	±18	100	D	***Raytheon device only Low noise audio amplifier***
uA739	1	300	NPN	10	1	86	±18	90	D	***Fairchild device only Low noise audio amplifier Needs frequency compensation***
LM381	**Not applicable**	**Not applicable**	NPN	15	—	112	±20	—	D	***Low noise amplifier Internally compensated***
CA3130	8	0.005	MOSFET	15	10	110	+16	90	S	***Ground sensing inputs Very high input impedance Needs frequency compensation***
CA3140	8	0.010	MOSFET	4.5	9	100	+36	90	S	***Ground sensing inputs Very high input impedence***
CA3160	6	0.005	MOSFET	4	10	110	+15	90	S	***Ground sensing inputs Very high / input impedence***
NE531 RC4531	2	400	NPN	10	35	96	±22	100	S	***Very fast op amp Needs frequency compensation***
CA3080	0.4	IABC 100	NPN	2	50	—	±18	110	S	***OTA device Programmable gain Current output***
CA3094	0.4	IABC 300	NPN	30	50	—	±12	110	S	***OTA device Programmable power switch/ amplifier***
TL080	15	0.4	JFET	3	13	83	±18	70	S	***JFET input op amps, with fast slew rate and wide bandwidth [TEXAS]*** — Pin for pin replacement for ***748***
TL081	15	0.4	JFET	3	13	83	±18	70	S	***741***
TL082	15	0.4	JFET	3	13	83	±18	70	D	***1458***
TL083	15	0.4	JFET	3	13	83	±18	70	D	***747***
TL084	15	0.4	JFET	3	13	83	±18	70	Q	***324***

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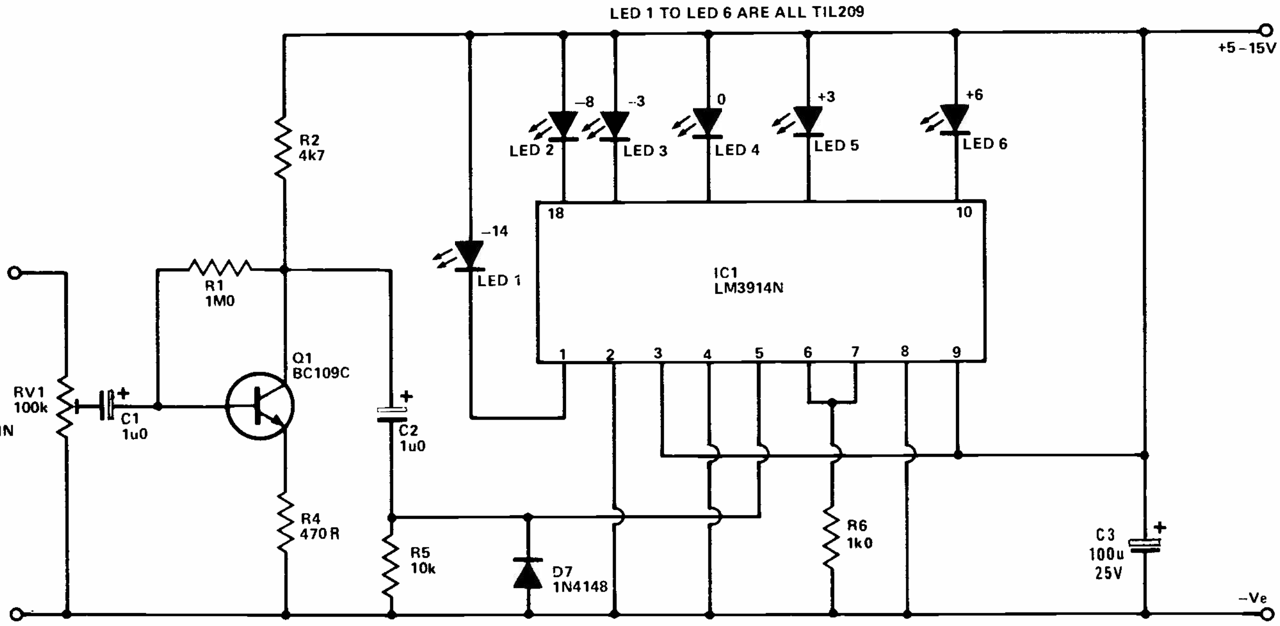
LED VU Meter

This simple peak reading VU meter circuit uses six LEDs to indicate six signal levels at − 14, − 8, − 3, 0, +3, and +6dB, or any other levels having the same spacing (e.g. − 17, − 11, − 6, − 3, 0, and +3dB, if preferred). About 24mV peak to peak is needed in order to activate the highest LED indicator, so the circuit is sufficiently sensitive to be used with any normal item of audio equipment.

The circuit is based on an LM3914N bargraph display driver device (IC1), which can be used to drive up to ten LEDs. This is connected so that with 0V12 at the input only the first LED indicator switches on. With the input raised to 0V24 the second LED switches on as well; 0V36 causes three LEDs to switch on and so on up to an input of 1V2 or more whereupon all ten LEDs are activated. In this circuit only LEDs 1, 2, 3, 5, 7 and 10 are included in the display, and these are D1 to D6 respectively.

The input signal is taken to a variable attenuator which enables the sensitivity of the circuit to be set at the correct level. The signal is then passed to a low gain common emitter amplifier based on Q1 which gives a tenfold boost in the sensitivity of the circuit. C2 couples the output from Q1 to the input of IC1. R5 is the input bias resistor for IC1, and D7 protects IC1 against an excess negative input voltage. R6 sets the current to each LED at about 12mA, but as IC1 responds only to positive half cycles the LEDs can switch on for a maximum of 50% of time. This gives an effective LED current of 6mA. The quiescent current consumption of the unit is about 8mA, rising to an absolute maximum of 44mA with all six LEDs activated.

To calibrate the unit, a 0dB test signal should be fed into the equipment and R1 adjusted for the lowest sensitivity that does not cause the 0dB LED to extinguish. The input impedance of the unit is about 80k and it will only lightly load the monitored equipment.

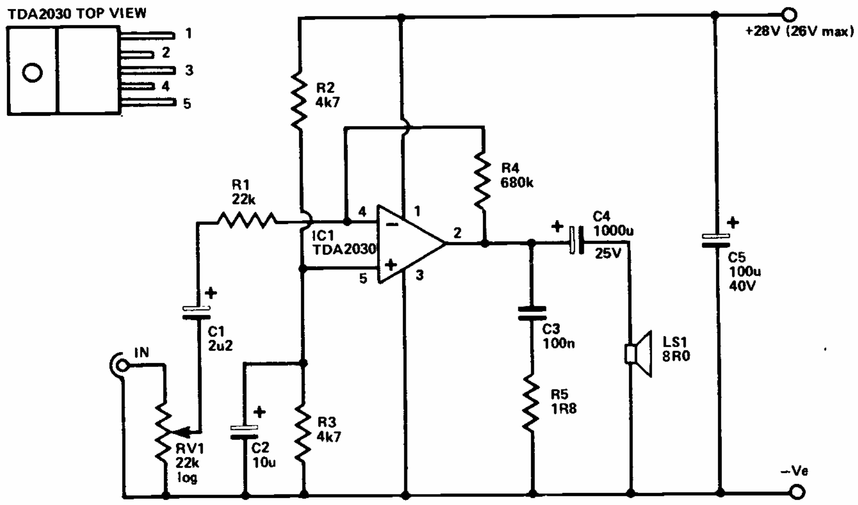


General Purpose Amplifier

This useful amplifier will provide an output power of up to about 8 watts rms at low distortion (less than 0.1% THD) into an 8 ohm loudspeaker when using a 28 volt supply. If used with a 4 ohm loudspeaker the output power is increased to about 12 watts rms or so, with the distortion being roughly doubled (although obviously still quite low). The circuit will operate with lower supply voltages, down to less than 9 volts, with a reduction in the maximum output power.

The circuit utilises a TDA2030 integrated circuit, a modern divice superior in performance and easier to use than most previous devices.

It is used in much the same way



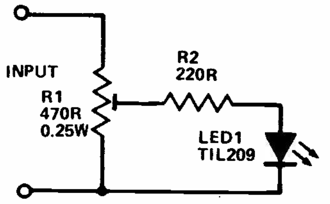
as an operational amplifier, and like an operational amplifier it has both inverting (−) and non-inverting (+) inputs. In this circuit it is used in the inverting amplifier mode.

The non-inverting input is biased to half the supply potential by R2 and R3; C2 decouples any hum or other noise from the supply lines. R4 biases the inverting input of IC1, with a negative feedback action stabilising the output at half the supply voltage under quiescent conditions. This ensures that the circuit can provide the highest possible unclipped peak to peak output voltage swing (and output power).

R4 also forms part of the network which sets the gain and input impedance of the circuit, with R1 forming the other half. The voltage gain is approximately equal to R4 divided by R1, about 30 with the specified values. This gives the circuit an input sensitivity of about 260mV rms for 8 watts into 8 ohms. The value of R1 sets the input impedance of the amplifier at 22k, but volume control VR1 shunts the input to give a minimum input impedance of 11kR.

Input and output DC blocking are given by C1 and C4 respectively. C3 and R5 are a Zobel network and aid the stability of the circuit. C5 provides supply decoupling and should be mounted close to IC1.

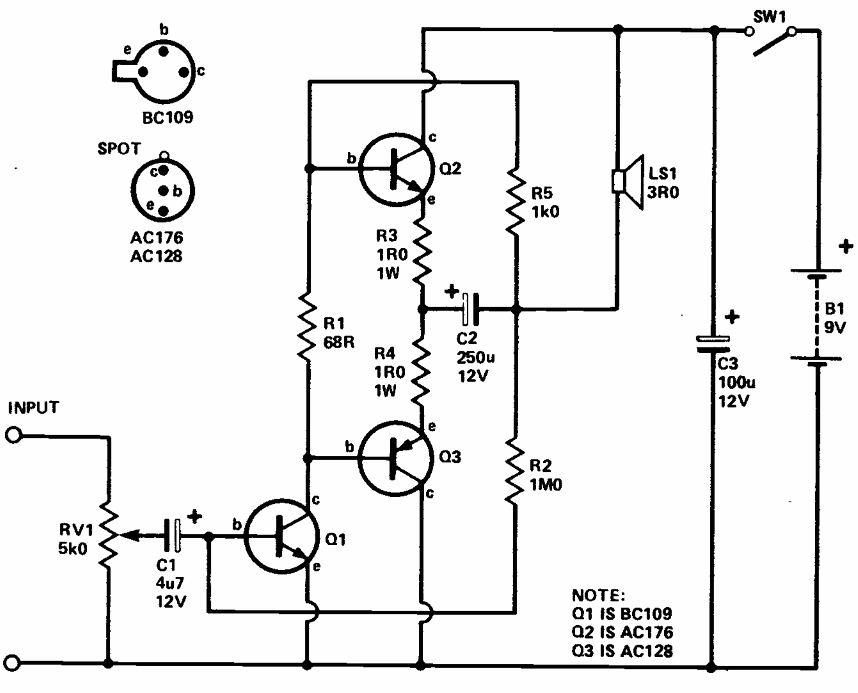
Clipping Monitor



WHEN using an amplifier at virtually its full output power there is a risk of "clipping" occurring. The amplifier is overdriven to the point where output peaks are flattened because the amplifier simply cannot provide a high enough output voltage. The distortion caused by clipping is often quite severe and readily apparent, but this is not always the case. It is not uncommon for the tweeters in loudspeakers to burn out due to overloading caused by clipping producing strong high frequency signals!

The circuit relies on the fact that with less than about 2V applied to a LED it will not pass any significant current and will fail to light up. If the voltage applied to the LED is only marginally increased above 2V, the LED avalanches, a heavy current flows and it glows brightly. R1 is adjusted so that the voltage applied to LED1 via current limiting resistor R2 is just sufficient to cause D1 to glow quite brightly when the amplifier is driven into clipping. With the amplifier driven just below the clipping threshold there is just enough voltage applied to the circuit to cause LED1 to visibly glow, although it does so at less than full brightness. If the amplifier is driven significantly below the clipping level there will be insufficient voltage fed to D1 to cause it to conduct and it will not light up.

Power Amplifier



Most Hi Fi amplifiers are similar to this design. The circuit uses a complementary output stage with one NPN and one PNP power transistor; this eliminates the need for an output transformer. Output power is close to 1W with reasonably low distortion.

The input signal is passed through the volume control RV1 and then via C1 to the base of Q1. The collector load for Q1 is composed of R1, R5 and the loudspeaker and the voltage at the collector will be about half the supply voltage, i.e. 4V5. The bases of Q2 and Q3 are also at the same voltage (very nearly) as the collector of Q1 because the value of R1 is so low.

At the junction of the emitters of Q2 and Q3 the voltage will also be very nearly 4V5. R3 and R4 limit the current through Q2 and Q3. When the amplified input signal is less than 4V5, Q2 is turned off but Q3 will conduct. When Q1 amplifies the signal to above 4V5 the reverse happens; Q2 conducts and Q3 is turned off.

The signals are combined at the common emitter junction of Q2 and Q3 and passed to the loudspeaker through the large electrolytic capacitor C2. Small values of C2 result in a poor low frequency response. Negative feedback is provided by R5 and R2; these ensure stability by reducing the gain slightly. R1 is included to provide a small amount of base bias for Q2 and Q3; more sophisticated designs use thermistors or diodes to prevent thermal runaway.

A disadvantage is the DC coupling of the transistors; the output pair should be 'matched pair', other types can be tried as long as they are also 'matched pairs'.

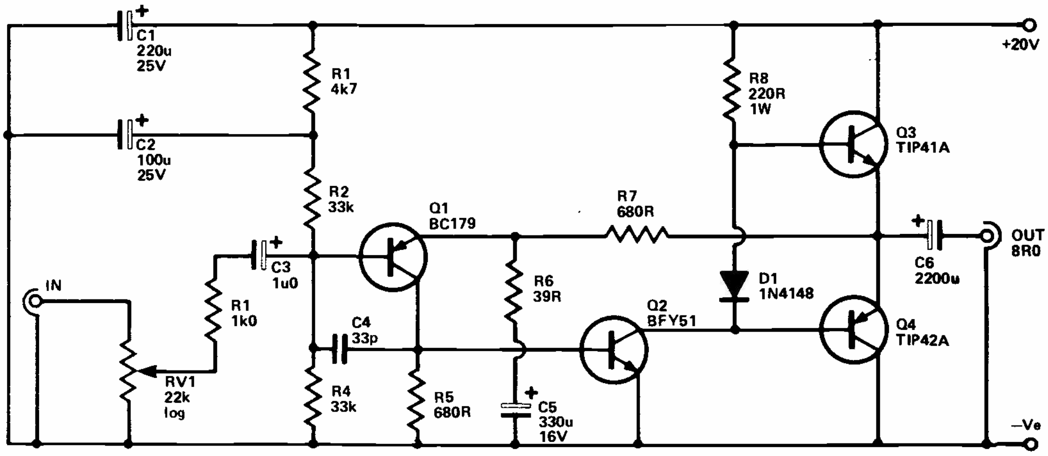
Cassette Radio Booster

This amplifier was designed as a booster to enable output powers of around 4 to 5 watts rms to be obtained from a radio/cassette unit. Of course, the amplifier is also suitable for other applications; it has an output sensitivity of approximately 350mV rms into 10k for maximum output, and is intended to feed an 8 ohm load.

The circuit uses a well known configuration which has common emiter input stage (Q1) direct coupled to common emitter driver stage (Q2), which is in turn direct coupled to the complementary emmitter follower output stage (Q3 – Q4). R7 provides virtually 100% negative feedback at DC, giving the circuit approximately unity voltage gain at DC. R1 R2 and R4 form a potential divider which bias the input of the amplifier to about half the supply potential and the output is also biased to about this level due to the DC unity gain. This bias level gives the optimum unclipped output voltage swing. R1, C1 and C2 filter out any hum or noise which might otherwide be coupled from the supply lines to the input via the bias circuit. R6 and C5 are used to decouple some of the feedback at audio frequencies, and thus give the unit a useful voltage gain at these frequencies.

D1 is used to give a small standing bias to the output transistors and, together with the fairly substantial amount of negative feedback, reduces cross-over distortion to an unnoticeable level. The emitter follower output stage gives the circuit a low output impedance so that the load can be efficiently driven with high output currents. Q3 drives the speaker during positive going output excursions while Q4 drives the speaker during negative output excursions. C6 provides DC blocking at the output, and C3 provides the same function at the input and C4 aids the stability of the circuit. RV1 is a volume control, and results will probably be best if the volume control on the cassette radio is set for a fairly high output (but not so high as to cause clipping), and the volume is adjusted using RV1.

The circuit requires a stabilised supply of 18 to 22 volts, and capable of providing up to 400mA. Q2 should be fitted with a clip-on TO-5 size heatsink. Q3 and Q4 are both fitted with commercially made, finned, bolt-on heatsinks.

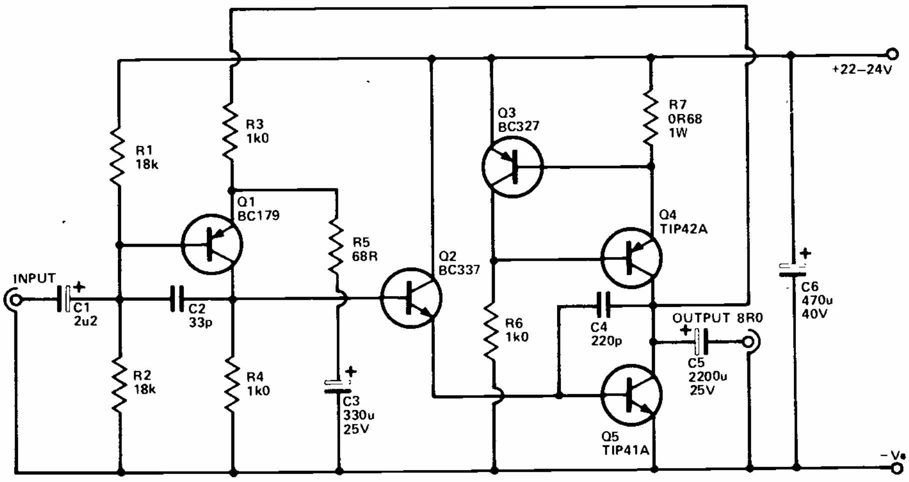


Class A Amplifier

This design gives an output power of about 4.1 watts rms into an 8 ohm load, but requires a supply of 22 to 24 volts at 1 amp. This gives an efficiency of about 19% at best, less than a third of the efficiency of many Class B designs.

Q1 is used in the common emitter input stage, and is direct coupled to the output stage via emitter follower buffer transistor, Q2.

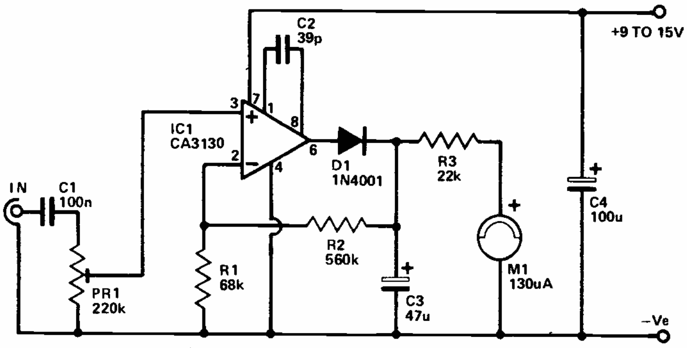
The latter is needed because of the fairly high drive current required by the output stage. Q5 is the output transistor, employed in the common emitter mode. It has a constant current source as its collector load, formed by Q3, Q4, and R7. The latter sets the output current of the circuit at just under 1 amp. The constant current generator load gives better efficiency than a load resistor and also gives good linearity.



R3 gives virtually 100% negative feedback over the amplifier at DC, giving unity voltage gain. By biasing the input to half the supply voltage using R1 and R2, the output is also biased to the required level of half the supply voltage. R5 and C3 decouple some of the feedback at audio frequencies, giving the circuit a sensitivity of about 380 mV rms for maximum output. C2 and C4 aid stability, while C1 and C5 provide DC blocking at the input and output respectively.

Q4 and Q5 must be mounted on a substantial heatsink.

Peak Reading VU Meter

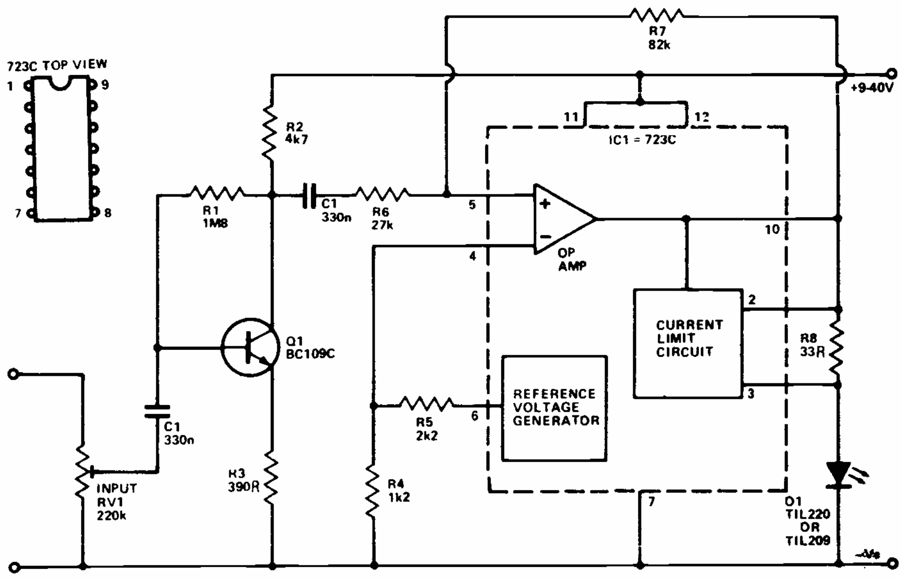


The type of VU meter normally employed in tape decks and other items of audio equipment is the average reading type. These can give misleading results on signals that have a pulse-like waveform of relatively low average amplitude for the peak amplitudes involved. This can lead to overloading and consequent distortion on signals of this type, eg piano and percussion. One way around this problem is to use a peak reading VU meter. This has a fast attack and slow decay time so that it responds properly to brief and intermittent signals. The normal response times for a unit of this type are 2.5ms attack and 1s decay. This unit roughly adheres to these figures.

IC1 is an operational amplifier which is used in the non-inverting mode. R1,2 form a negative feedback network which sets the closed loop voltage gain of the circuit at a little under ten. D1 is included at the output so that IC1 can supply an output current, but a current cannot flow into the output of IC1. The feedback is taken from the junction of D1, R2 etc., so that the input voltage appears here amplified by about ten times and the feedback overcomes the non-linearity of D1. C3 is rapidly charged to the peak output voltage as it is fed from the fairly low impedance of IC1 and D1. Its only discharge paths are through the much higher impedances of R1-R2 and R3-M1. This gives the circuit the required fast attack and slow decay times. M1 responds to the voltage across C3, which is, of course, proportional to the peak positive input level (the circuit is a halfwave type and does not respond to negative going inputs). The VU meter movement used in the prototype had a FSD value of 130uA, but the circuit should work with any type having a sensitivity of between about 50 and 200uA.

R1 biases the non-inverting input of IC1 to the negative rail and also enables the sensitivity of the circuit to be djusted to the correct level. At maximum sensitivity, less than 1V peak to peak is needed for FSD of M1. Current consumption is only 400uA.

Peak Level Indicator



Peak audio level indicators can be used in tape recorders, amplifiers, mixers, and other radio equipment to provide a visual overload warning, and unlike slower responding VU meters, they produce a proper response to fast transients. The circuit is based on the inexpensive 723C device which, although primarily intended for use as a voltage regulator, can be adapted to work well in many other applications.

The 723C has a highly stable 7V (nominal) reference voltage available at pin 6. This is coupled to the inverting input of an operational amplifier (which is also part of the 723C device) via an attenuator, R4,5. This gives a stable reference potential of a little over 2V at the inverting input. The input signal is coupled by way of sensitivity control RV1 to a common emitter amplifier based on Q1 and fed to the non-inverting input of the operational amplifier by C2 and R6. Under quiescent condition or with a negative going signal at Q1 collector, the non-inverting input will be at a lower potential than the inverting one, and the output of the amplifier will be low. If a positive going signal reaches a high enough amplitude, though, the non-inverting input will reach a higher potential than the inverting one causing the output to go high. D1 is then switched on with a current that is determined by the output of the amplifier and which is largely independent of the supply voltage. Discrete resistor R8 actually sets the output current. The specified value gives a nominal 20mA LED current. R7 provides positive feedback which ensures that D1 is either fully on or off. It also tends to hold D1 in the on state for slightly longer than would otherwise be the case, thus giving a clearer indication of a brief overload.

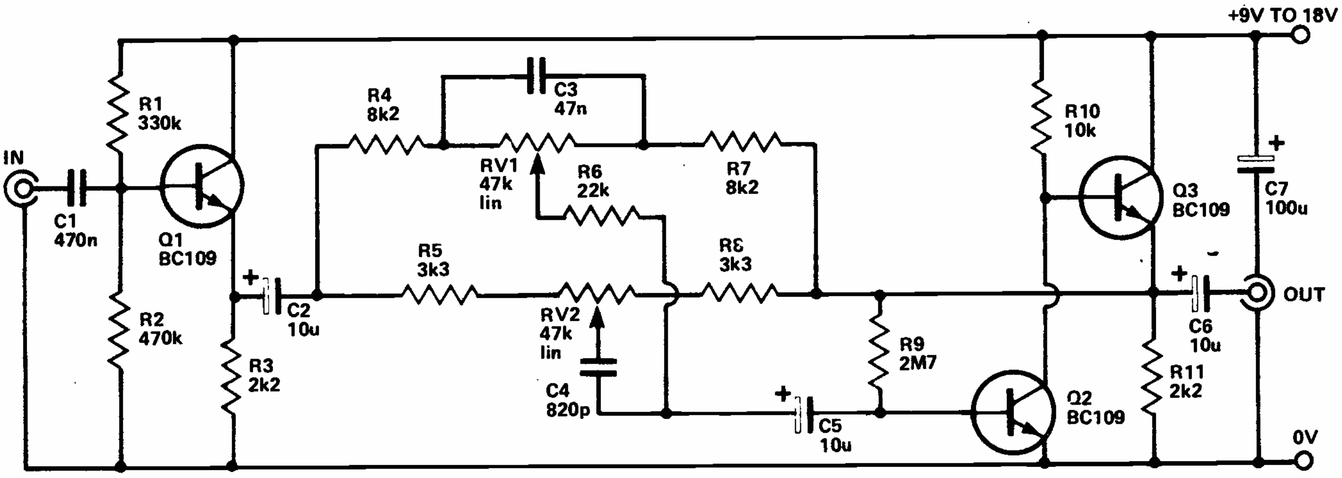
The unit can be adjusted to respond to input levels down to about 100mV rms, which should be more than adequate for all normal requirements. RV1 is adjusted for the lowest sensitivity that causes D1 to come on with an input signal level equal to the lowest overload level. Quiescent current consumption is about 4mA.

Active Tone Controls

This tone control circuit is easy to incorporate in a stereo amplifier, disco unit or whatever, as it has a high input impedance (over 100kR), a nominal voltage gain of unity and a low output impedence. The usual bass and treble controls are included in the unit, with about 12 dB of boost and cut being available at 100Hz and 10kHz. The noise and distortion produced by the circuit are both extremely low due to the large amount of negative feedback used and the unit can handle output signal levels of several volts rms without clipping.

Q1 is used in a straightforward emitter follower buffer stage that gives the unit a high input impedance. C2 couples the output of Q1 into the tone control circuitry. This is an active circuit which provides frequency-selective negative feedback over an amplifier. The amplifier uses Q2 as a conventional common emitter stage direct-coupled to emitter follower output transistor Q3. The latter gives the unit a low output impedance.

The tone control networks are slightly simpler than the usual Baxandall configuration, but give a perfectly acceptable level of performance. RV1 controls the bass while RV2 is the treble control. Feedback is at a maximum with the sliders of the potentiometers to the right and at a minimum with the sliders set fully to the left. Of course the gain of the circuit is inversely proportional to the level of feedback. Maximum feedback therefore corresponds to maximum cut and not to full boost. The current consumption of the circuit is a little under 1mA per supply volt.

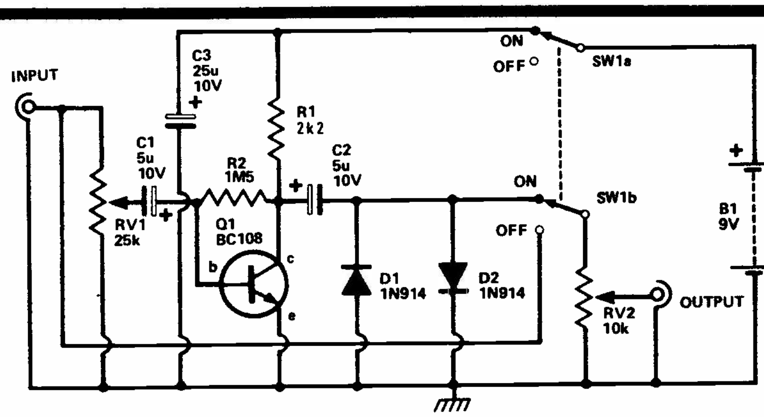


Peak Limiter

The circuit is designed to take almost any audio input but the output will have all peaks above a certain level, which can be adjusted, eliminated.

The circuit can either be wired into a receiver circuit or directly from a headphone socket. If wired into circuit permanently, RV1 should take the place of the normal volume control and the output should be wired to the point which was previously connected to the volume control slider. An extra control plus a switch will also have to be mounted on the receiver front panel. The circuit can either be left in permanently,as at most settings it will not affect the signal, or it can be switched.

The output of the receiver is taken to the input and amplified by Q1 which is connected in the common emitter mode. This transistor will considerably increase the audio level and this is applied via a DC blocking capacitor, C2 to the two silicon diodes D1 and D2. In the normal way these diodes will not have any bias voltage applied across them and so they will present a high resistance, and will not affect the output in any way. However, as soon as the output from the amplifier exceeds about 0V6, the diodes will conduct and short the output to the negative line. Two diodes are needed, one connected each way around so that both positive and negative going peaks are shorted out. The idea is to make sure that whatever the input level across RV1, it can be amplified so that at least 0V6 can be applied across the diodes. Since RV1 is adjusted so that the level is always the same a volume control has been included in the circuit so that the output level can be controlled in the usual way; this is accomplished by RV2.



To limit the noise, the input level is increased until the signal is just distorting, then backed off slightly so that no distortion is heard on the peaks. RV2 is then adjusted as a normal volume control. If RV1 is adjusted well below the limiting level and RV2 is adjusted for normal listening levels, the circuit has no effect. However, it is a simple matter to include SW1 which will bypass the circuit. The supply voltage can be taken from a battery as shown in the circuit, the current drain being very small, or from the receiver's supply.

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AY5 1224A 240p
AY5 1315 600p
AY5 4007D 520p
CA3028A 120p
CA3019 80p
CA3046 70p
CA3048 225p
CA3059 300p
CA3080E 72p
CA3086 48p
CA3089E 225p
CA3090AO 375p
CA3130E 90p
CA3140E 50p
CA3160E 100p
CA3161E 140p
CA3162E 450p
CA3189E 300p
CA3240E 120p
CA3280G 200p
DAC1408-8 200p
HA1388 270p
ICL7106 850p
ICL8038 300p
ICM7555 80p
LC7120 600p
LC7130 475p
LC9109 1000p
LF347 180p
LF351 48p
LF353 100p
LF356P 95p
LF357 120p
LM10C 425p
LM301A 27p
LM311 75p
LM318 200p
LM319 225p
LM324 45p
LM335Z 140p
LM339 65p
LM348 75p
LM358P 75p
LM377 175p
LM380 75p
LM381AN 180p
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LM389 95p
LM393 100p
LM394 300p
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LM710 50p
LM725 350p
LM733 100p
LM741 18p
LM747 70p
LM748 35p
LM2917 200p
LM3302 140p
LM3900 65p
LM3909 95p
LM3911 130p
LM3914 210p
LM3915 225p
LM3916 225p
LM13600 125p
M51513L 300p
M51516L 500p
MB3712 250p
MC1310P 150p
MC1458 40p
MC1495L 350p
MC1496 70p
MC3340P 120p
MC3403 120p
MK5039B 750p
ML920 800p
MM57160 820p
NE531 250p
NE555 20p
NE558 50p
NE564 420p
NE565 130p
NE566 165p
NE567 140p
NE570 425p
NE571 425p
NE5534A 250p
PLL02A 500p
RC4136 70p
RC4151 200p
S566B 260p
SAD1024A 1250p
SFF96364 800p
SL490 350p
SN76477 175p
SP8515 750p
TA7205 250p
TA7120 200p
TA7204 250p
TA7222 200p
TA7310 200p
TAA621 275p
TBA641BX1 300p
TBA651 200p
TBA800 90p
TBA810 100p
TBA820 80p
TBA950 300p
TCA210 350p
TCA220 350p
TCA940 175p
TDA1004A 300p
TDA1008 320p
TDA1010 225p
TDA1022 600p
TDA1024 120p
TDA1034B 250p
TDA1170 300p
TDA2002V 325p
TDA2020 320p
TL071/81 45p
TL072/82 75p
TL074 130p
TL084 110p
TL094 200p
TL170 60p
TL430C 70p
UAA170 170p
UA2240 300p
UDN6118 320p
UDN6184 320p
ULN2003 100p
UPC575 400p
UPC592H 200p
UPC1156H 300p
XR2206 300p
XR2207 400p
XR2211 600p
XR2215 675p
ZN414 90p
ZN419C 225p
ZN423E 150p
ZN424E 135p
ZN425E 360p
ZN427E 625p
ZN1034E 200p

TRANSISTORS
AD161/2 45p
BC107/8 11p
BC109 11p
BC117 20p
BC147/8 9p
BC149 10p
BC157/8 10p
BC159 11p
BC169C 12p
BC172 12p
BC177/8 17p
BC179 18p
BC182/3 10p
BC184 11p
BC187 30p
BC212/3 11p
BC214 12p
BC237 15p
BC327 16p
BC337 16p
BC338 16p
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BC548C 9p
BC549C 18p
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BC559C 18p
BCY70 18p
BCY71/2 22p
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BD135/6 64p
BD139 58p
BD140 60p
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BD235 85p
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BF256B 70p
BF257/8 32p
BF259 36p
BFR39 25p
BFR40/1 25p
BFR79 25p
BFR80/1 25p
BFX29 40p
BFX30 34p
BFX84/5 40p
BFX86/7 30p
BFX88 30p
BFY50 30p
BFY51/2 30p
BFY56 33p
BFY90 80p
BRY39 45p
BSX19/20 24p
BU104 225p
BU105 190p
BU108 250p
BU109 225p
BU126 150p
BU180A 120p
BU205 200p
BU208 200p
BU406 145p
BUY69C 350p
E310 50p
MJ2501 225p
MJ2955 90p
MJ3001 225p
MJE340 60p
MJE2955 100p
MJE3055 70p
MPF102 45p
MPF103/4 40p
MPF105 40p
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MPSA12 50p
MPSA13 50p
MPSA20 50p
MPSA42 50p
MPSA43 50p
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TIP33A 90p
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TIP34C 160p
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TIP36C 340p
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TIP142 130p
TIP147 130p
TIP2955 78p
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ZTX500 15p
ZTX502 18p
ZTX504 30p
VN66 80p
2N697 25p
2N698 45p
2N706A 30p
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2N918 45p
2N930 18p
2N1131/2 36p
2N1613 25p
2N1711 25p
2N2102 70p
2N2160 350p
2N2219A 30p
2N2222A 30p
2N2369A 25p
2N2484 30p
2N2648 45p
2N2904/5 30p
2N2906A 30p
2N2907A 30p
2N2926 9p
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2N3773 300p
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2N3823 70p
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2N4061/2 18p
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2N4125/6 27p
2N4401/3 27p
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2N4871 60p
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2N5089 27p
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2N5457/8 40p
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2N5460 60p
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2N6027 48p
2N6052 300p
2N6059 325p
2N6107 65p
2N6247 190p
2N6254 130p
2N6290 65p
2SC1172 150p
2SC1306 150p
2SC1307 220p
2SC1957 90p
2SC1969 195p
2SC2028 120p
2SC2029 250p
2SC2078 200p
3N128 120p
3N140 120p
3N141 110p
3N201 110p
3N204 120p
40290 260p
40361/2 75p
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40409 100p
40410 100p
40411 300p
40594 120p
40595 120p
40673 75p
40871/2 100p

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BY127 12p
BYX36-300 20p
OA47 8p
OA90/91 9p
OA95 9p
OA200 9p
OA202 10p
1N914 4p
1N916 7p
1N4148 4p
1N4001/2 5p
1N4003/4 6p
1N4005 6p
1N4006/7 7p
1N5401/3 14p
1N5404/7 19p
IS920 9p

HEATSINKS
For TO220 Voltage Regs. and transistors 22p
For TO5 12p

BRIDGE RECTIFIERS
1A 50V 19p
1A 100V 20p
1A 400V 25p
1A 600V 30p
2A 50V 30p
2A 100V 35p
2A 400V 45p
3A 200V 60p
3A 600V 72p
4A 100V 95p
4A 400V 100p
6A 50V 80p
6A 100V 100p
6A 400V 120p
10A 400V 200p
25A 400V 400p

ZENERS
2.7V-33V
400mW 9p
1W 15p

TRIACS PLASTIC
3A 400V 60p
6A 400V 70p
6A 500V 88p
8A 400V 75p
8A 500V 95p
12A 400V 85p
12A 500V 105p
16A 400V 110p
16A 500V 130p
T2800D 130p

THYRISTORS
1A 50V 70p
1A 400V 90p
3A 400V 100p
8A 600V 140p
12A 400V 160p
16A 100V 180p
16A 400V 180p
BT106 110p
C106D 45p
MCR101 36p
TIC44 27p
2N3525 130p
2N4444 140p
2N5060 34p
2N5064 40p

LOUD-SPEAKERS
Size
2½" 64R 80p
2½" 8R 80p
2" 8R 90p
1½" 8R 100p

VEROBOARDS
0.1 (Copper clad)
2.5 x 3.75" 75p
2.5 x 5" 85p
3.75 x 3.75" 85p
3.75 x 5" 95p
3.75 x 17" 340p
4.75 x 17.9" 420p
Pkt of 100 pins 50p
Spot face cutter 86p
Pin insertion tool 110p
Vero Wiring Pen 340p

ANTEX
Irons
C 15W 450p
CX 17W 460p
CCN 15W 470p
X25 470p

SPARE BITS
C/CX/CCN 55p
X25 55p

SPARE ELEMENTS
C/CX/X25 180p
CCN 200p

RELAYS
PCB mounting
6V Coil OUA SPDT 2A at 24V DC 180p
12V Coil OUA SPDT 2A at 24V DC 180p
12V Coil OMI DPDT 5A at 24V DC 240V AC 200p
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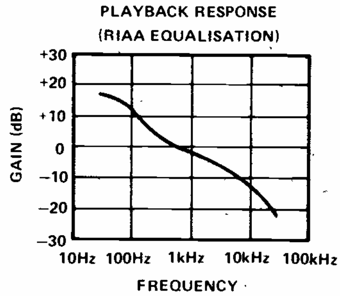
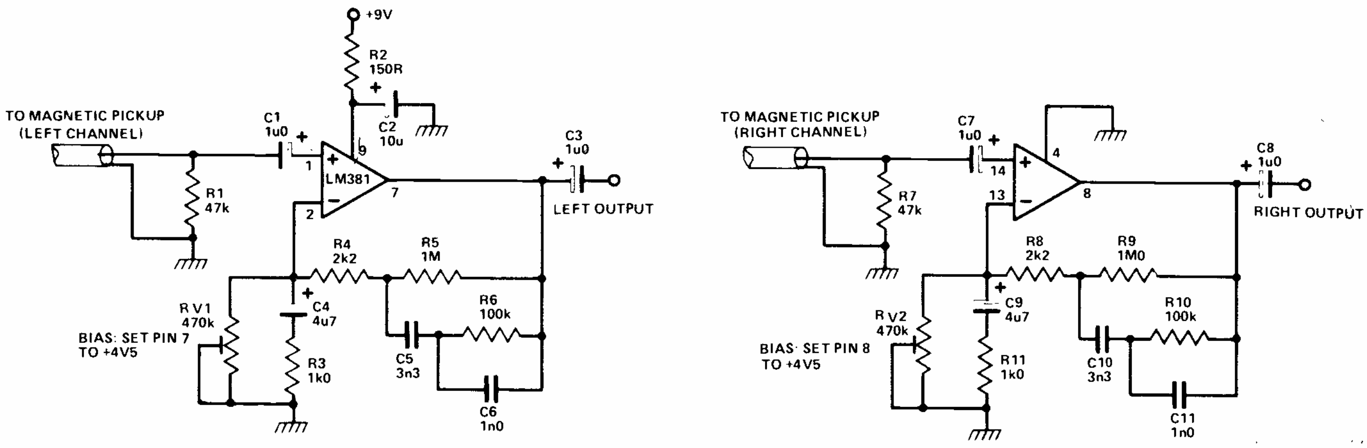
RIAA Stereo Preamplifier

Records are cut with a frequency response such that when they are replayed with a magnetic pickup and a preamplifier with RIAA equalisation (Recording Industry Association of America) the reproduced sound will be as similar to the original as possible.

The disc is cut at constant amplitude, except from 500Hz to 2120Hz where it is cut at constant velocity. When this disc is replayed with a magnetic pickup, the relative output voltage rises with frequency, due to the fact that the magnetically generated voltage is proportional to the velocity of the stylus as it moves sideways in the groove. To restore the original sound quality, a preamplifier with a frequency response that gives decreasing output with increasing frequency is required. This response curve is known as the RIAA equalisation and it is tailored accurately to fit the cutting and replay processes. The signal level from a magnetic pickup is low, generally 20mVpp and so a low noise preamplifier is needed.

The circuit shows a realisation of this requirement. The low noise amplifier is the LM381 made by National Semiconductors. A DC bias control is included (RV1, RV2), and the feedback components generate the RIAA curve. Use screened cable for the wiring to the pickup, keep the circuit away from transformers (and the pickup and its wiring) and connect all the earths together, near to the IC.



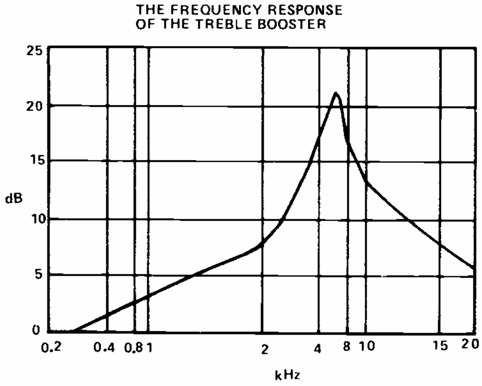


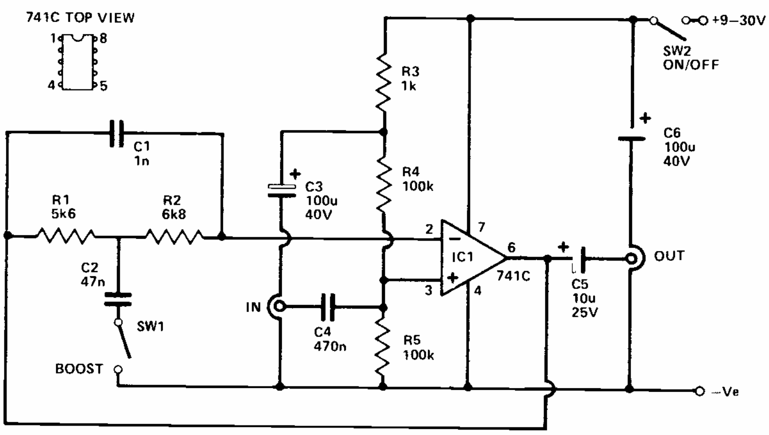
Treble Booster

A treble booster circuit can be used with an electric guitar (and also electronic instruments) to boost the higher order harmonics and give a more brilliant sound. A circuit of this type gives a fairly flat response at bass and most middle audio frequencies, with the upper-middle and lower treble frequencies being given a substantial amount of boost. It is normal to use only a modest amount of emphasis to the upper-treble in order to give good stability and a low noise level, and this also prevents the output from sounding too harsh. The frequency response is shown in the accompanying graph.

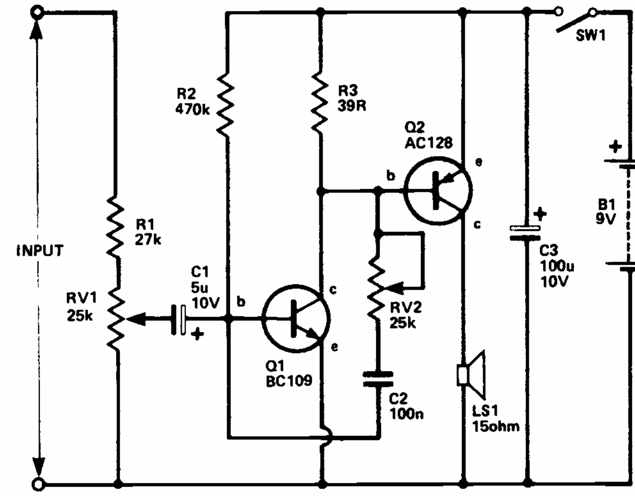
The circuit is basically an op amp used in the non-inverting amplifier mode. The non-inverting input is biased by R4 and R5 via a decoupling network which is comprised of R3 and C3. C4 and C5 give DC blocking at the input and output respectively. With SW1 open there is virtually 100% negative feedback through R1, R2 and C1, giving the circuit unit gain and a flat response. Closing SW1 brings C2 into circuit, and this decouples some of the feedback through R1 and R2 at frequencies of more than a few hundred Hz, giving the required rising response. Feedback through C1 at high treble frequencies causes the response to fall away above about 5k5Hz, and prevents the very high frequency harmonics from being excessively emphasised.

THE FREQUENCY RESPONSE OF THE TREBLE BOOSTER





Simple Amplifier



The term "amplifier" covers a very wide range, from a one transistor preamp to an ultra sophisticated high power hi-fi system. There is little doubt that the circuit shown here is very simple. The output is in the order 250mW – quite sufficient for most purposes and comparable to that of the average transistor radio. The distortion level is rather high, being about 5%.

The amplifier is also reasonably sensitive and will give full output with an input of about 50mV. Input impedance is about 50kR.

The slider from the volume control is connected to the base of Q1 via a DC blocking capacitor. Q1 is connected as a conventional common emitter amplifier with R2 providing the base bias and R3 acting as the collector load. This stage is directly connected to the second transistor which is a PNP type. In this way the current passing through Q1 provides the bias for the second transistor. The output of the second transistor is connected directly to the speech coil of the loudspeaker. This is not normally good practice since the standing current in the output transistor continually biases the coil either slightly in or out from its usual operating point. However if a large speaker is used, as it should be, this has very little effect and, since we are not aiming at hi-fi, it does not matter.

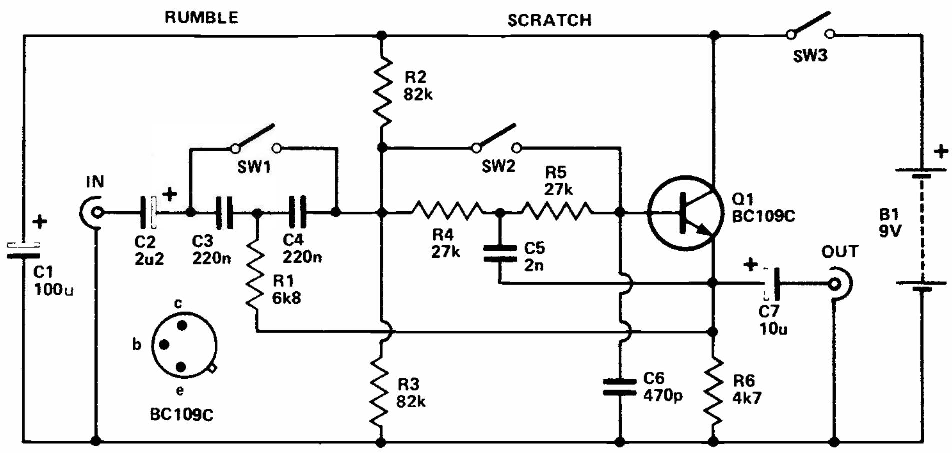
The tone control comprises C2 and RV2 which are connected between the collector and base of Q1. At high resistance settings RV2 has little effect but on minimum settings the 100nF feeds back the high frequencies out of phase, thus cancelling them.

For this circuit to work properly, R3 must be selected with great care. The value shown here of 39 ohms is a typical one and, although it may be used for initial setting up to ensure the circuit is operating, the value should be found by experiment. If it is too low there will be severe distortion at higher volume settings. If it is too high the current drain will be excessive even though the quality of reproduction will be good.

It is very important that Q2 is fitted with a heatsink as it will get very hot.

The speaker impedance is not all that critical and in the prototype speakers with an impedance as low as 8 ohms and as high as 80 ohms all worked well, although changing the speaker impedance will also necessitate a change in the value of R3.

Scratch And Rumble Filter



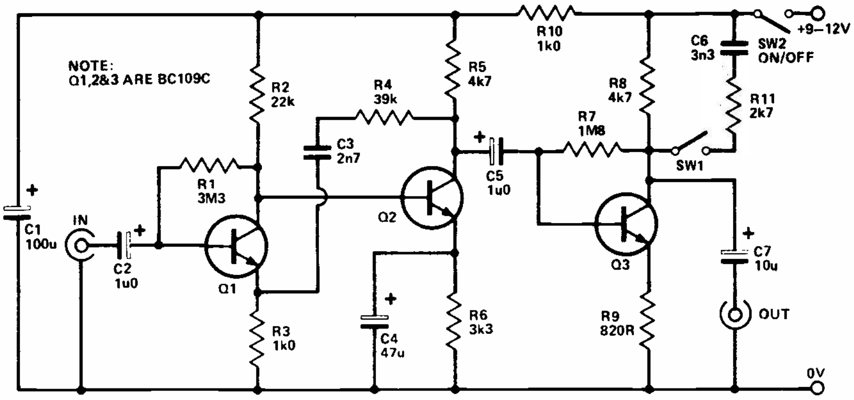
This is a 12 dB per octave add-on scratch and rumble filter which can be connected into the 'tape Monitor' or some similar facility of the amplifier.

It is a conventional second order filter circuit having passive high pass filter formed by the series capacitance C3 and C4, plus the parallel resistance of R2 and R3 (the latter also being used to bias emitter follower transistor Q1). A passive filter of this type gives only a very slow initial roll off, and an ultimate attenuation rate of only 6 dB per octave. A bootstrapping resistor is therefore used to improve performance. Above the cut-off frequency, where the gain of the circuit would otherwise fall off somewhat, R1 has the effect of reinforcing the input signal. Well below the cut off frequency, losses through C4 result in the signal level at Q1 emitter being well below that at the junction of C3 and C4. This results in some of the signal at the junction of C3 and C4 being tapped off through R1, with C3 and R1 effectively forming a second high pass filter network. This eliminates the slow initial roll off rate (in fact there is a small and insignificant peak of about 0.5dB above the cut off frequency) and speeds up the attenuation rate to a nominal 12dB per octave.

The low pass filter works in much the same way as the high pass one, except of course, the R and C filter elements have been transposed so as to give the correct filter action.

With the specified component values the rumble filter response falls below unity at approximately 45Hz, reaches the -6 dB point just above 30Hz, and then falls away at a nominal 12dB per octave. The scratch filter response crosses the unity gain point at about 6k5Hz, reaches the -6dB point at approximately 10kHz, and then falls away at a nominal 12dB per octave.

Cassette Preamplifier



Used in conjunction with one of the cassette mechanisms currently available on the surplus market (or a mechanism removed from an old recorder or player) this preamplifier circuit makes an inexpensive but useful cassette player for use with a hi-fi system.

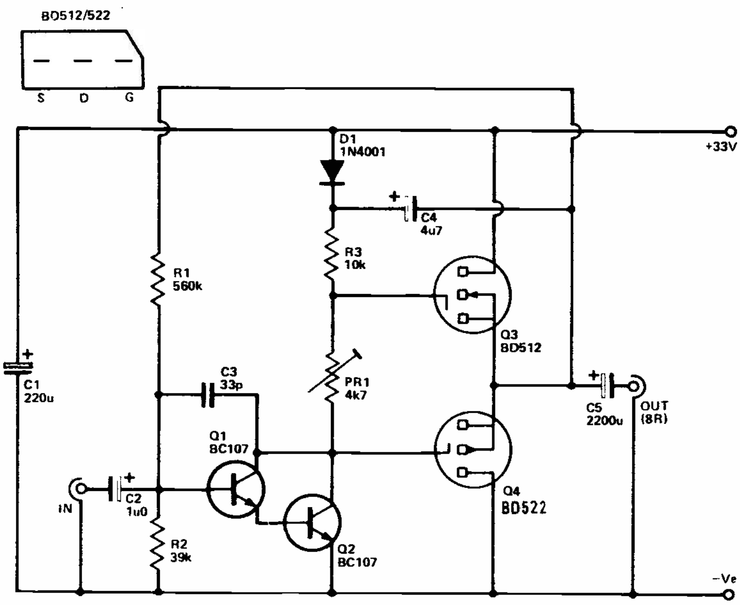
The output signal level from a cassette tape head is typically about 500µV or so at middle audio frequencies for a mono head and about half this level for a stereo type. The preamplifier must, therefore, provide a considerable amount of voltage gain in order to match this to a hi-fi amplifier, since these require a signal level about 1,000 times higher. It is also necessary for the preamplifier to provide equalisation, because the output from a tape head rises at a rate of 6dB per octave. However at higher audio frequencies, tape heads are not very efficient and require a much less rolloff.

Q1 and Q2 are used in a conventional two stage, direct coupled, common emitter amplifier and the frequency-selective negative feedback through C3 and R4 provides the appropriate equalisation. These also set the midband voltage gain of the input stage at about 46dB. With such a low input level it is obviously necessary to use low noise transistors (Such as the BC109C) in order to obtain good results. Running Q1 at a low collector current, about 200µA, also helps.

Q3 is used as a low gain common emitter stage, which provides the additional amplification. R9 introduces negative feedback, which controls the voltage gain of Q3 and the specified value gives a gain of about 14dB. For a stereo unit R9 should be reduced to 390R in order to give increased gain, to compensate for the lower output of a stereo tape head.

When playing a Dolby B encoded cassette SW1 can be closed; this gives a small degree of treble cut which provides a reasonably flat overall response.

VMOS 10 Watt Amplifier



At first sight this circuit may seem to be a straightforward Class B design having an emitter follower, complementary output pair and Darlington Pair common emitter drive stage. However, the output devides are, in fact, complementary VMOS transistors used in the source follower mode (the FET equivalent of the emitter follower).

R1 and R2 are used to bias the unit to give the optimum quiescent output potential and they provide overall negative feedback, which improves the quality of reproduction. D1 and C4 are bootstrapping components, enabling the gate drive voltage to Q3 to go above the positive supply potential, giving improved efficiency to the circuit. R3 is the main collector load for Q2 and PR1 is used to give a standing bias on the output transistors that gives a quiescent current consumption of about 25mA. The thermal compensation circuitry normally used is totally unnecessary in this circuit, since VMOS devices do not suffer from thermal runaway. In fact the quiescent bias current will drop slightly as the output devices heat up, but not sufficiently to give rise to significant crossover distortion.

C2 and C5 provide DC blocking at the input and output respectively, while C1 is a supply decoupling component. C3 gives a degree of high frequency attenuation and aids the stability of the circuit.

Although the current in the driver stage, only about 1mA, may seem to be totally inadequate, it is in fact more than sufficient since the VMOS devices have extremely high input impedances and consume no significant input current. This is one of their main advantages over bipolar devices. One disadvantage in this particular application is lower efficiency due to the higher threshold voltages and on resistance of VMOS transistors in comparison to bipolar devices. However, the circuit will give an output of 10W rms using a supply voltrage of about 33V or so (with a current drain of up to about 600mA). An input of about 500mV rms is needed for maximum output.

Note: The output devices do not have internal zener protection diodes and the appropriate handling precautions should be taken.

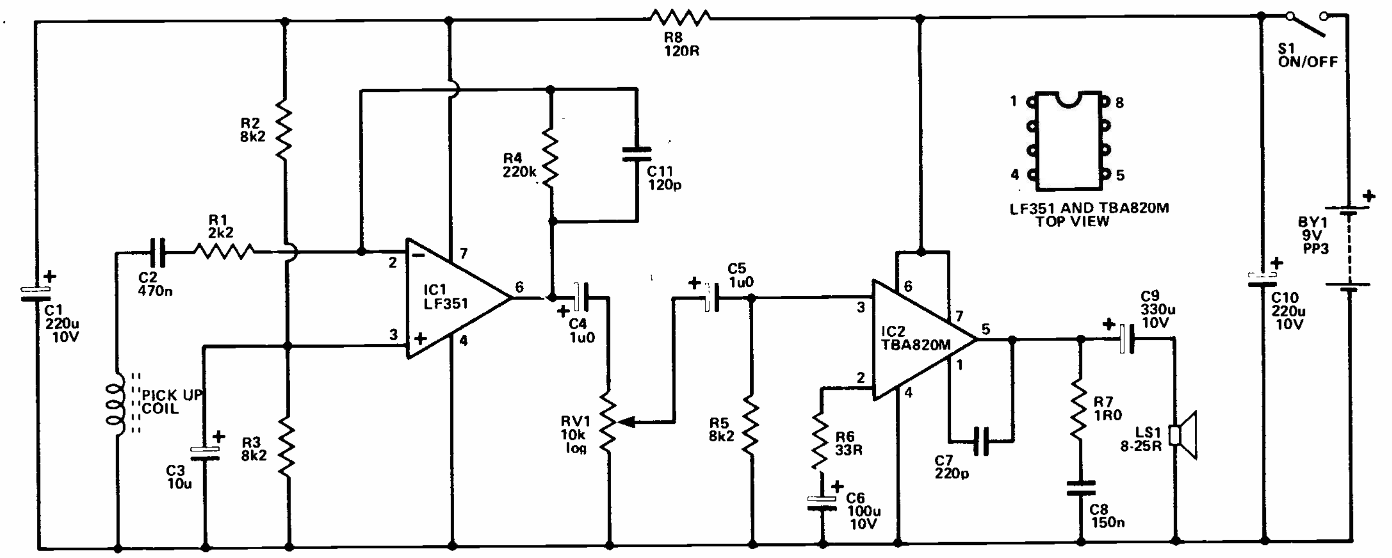
Telephone Amplifier

A telephone amplifier enables more than one person to follow a telephone conversation. The unit described here, in common with all normal units of this type, requires no direct connection to the telephone. Instead, the special pick-up coil has a built-in rubber suction cap that enables it to be easily attached to the telephone base. This produces a very weak signal from the magnetic field radiated by an inductive component inside the telephone, but satisfactory results can be obtained if it is fed to low noise, high gain amplifier. It would of course be possible to use a much simpler circuit if a direct connection to the telephone were to be made, but this would make installation more difficult and it is ILLEGAL to make a direct connection to a Post Office telephone anyway.

The preamplifier stage of the unit is based on IC1 which is a low noise op amp having a FET input stage. This is used in the conventional inverting audio amplifier mode and the negative feedback network, R1, 4, sets the voltage gain at about 40dB. (100 times). C11 reduces the gain slightly at high frequencies in order to obtain an improved signal to noise ratio.

C4 couples the output from the preamplifier to volume control, RV1, and from here the signal is coupled to the power amplifier by C5. The output stage uses the TBA820M, a class B amplifier which will give an output power of a few hundred milliwatts rms. The closed loop voltage gain of the device is determined by the value of R6, about 25dB. (180 times) with the specified value. This gives the required very high overall gain in conjunction with the preamplifier's gain. C7, R7 and C8 are needed in order to maintain stability.

The quiescent current consumption of the unit is only about 5mA, but this rises to as much as 50mA or so at high volume levels. The best position for the pick-up coil on the telephone base (not the handset) can be located with a little experimentation.



Stereo Synthesiser

There are two common methods of producing a pseudo stereo effect from a mono signal; playing the mono signal from the two speakers in antiphase, and the use of frequency selective techniques which normally consists of directing lower frequency signals into one channel and higher frequency signals into the other. This circuit uses the second technique, but can additionally give antiphase signals which can give a better effect, especially when using headphones.

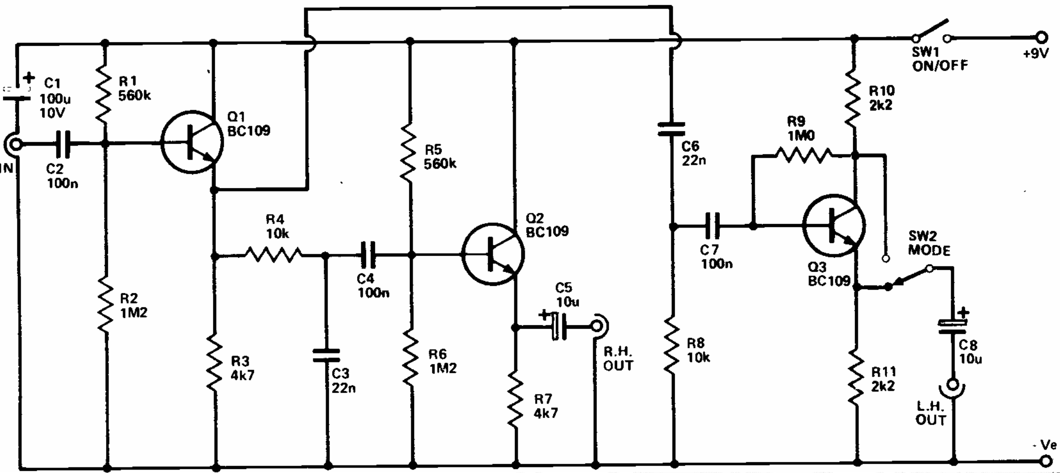
Q1 is used as an emitter follower buffer stage which ensures that the two filter networks fed from its output are driven from a low impedance source. If these were driven direct from the input, it is quite possible that they would be fed from a source impedance of a few kilohms or more, which would be quite sufficient to alter their effective characteristics.

The two filters are formed by R4 and C3 (low pass), and C6 plus R8 (high pass). A high roll off rate is by no means essential in this application and the 6dB per octave attentuation rate of simple RC filters such as these is perfectly adequate. The -3dB point of each filter is at approximately 800Hz and the combined output of the filters, therefore, gives a virtually flat response with no significant peaks or troughs.

Q2 is connected as an emitter follower buffer stage and this ensures that there is minimal loading on the low pass filter. Q3 similarly ensures that there is minimal loading on the high pass filter, but this device is also used as a phase splitter. With SW2 switched to take the output from Q3's emitter, Q3 effectively operates as an emitter follower and gives no phase inversion. With SW2 switched to take the output from Q3's collector, Q3 then effectively acts as a common emitter stage with 100% negative feedvack (and unity voltage gain) due to R11. It also provides a 180° phase shift so that the two output signals are in antiphase. An in-phase relationship is needed to give a good central stereo image and the use of antiphase signals tends to give an impression of increased channel separation.

In a stereo orchestral recording, it is normal for the violins to come from the left hand channel, with the cellos and basses from the right hand channel.Therefore, the high frequency signals are fed to the left channel and the low frequency signals are fed to the right channel so that the unit provides a similar effect (although it will obviously function properly with the outputs connected either way).

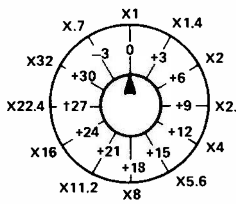
Stereo Synthesiser



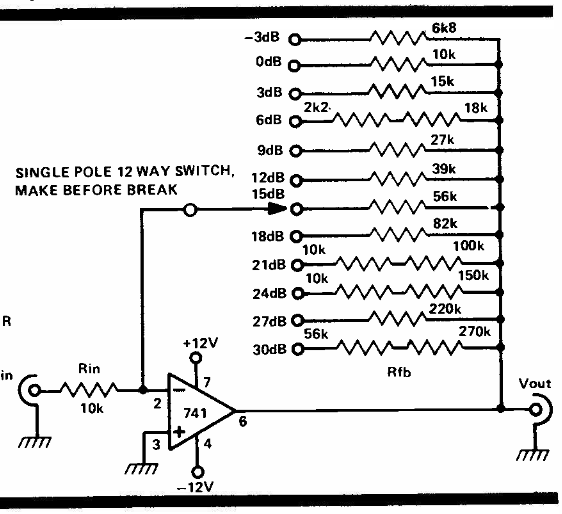
Preamp With dB-Stepped Gain Control

A handy little piece of test equipment is a preamplifier with stepped gain control selected by a rotary switch. The circuit here uses a single IC, 14 resistors and a single-pole, 12-way rotary switch.

The voltage gain of an op-amp is determined by the ratio of R_{fb}/R_{in}; thus by having R_{fb} switched, the voltage gain can be varied. The input impedance of the preamplifier is set by R_{in} to 10k.



OUTER RING – VOLTAGE GAIN AS A MULTIPLIER
INNER RING – VOLTAGE GAIN IN dB



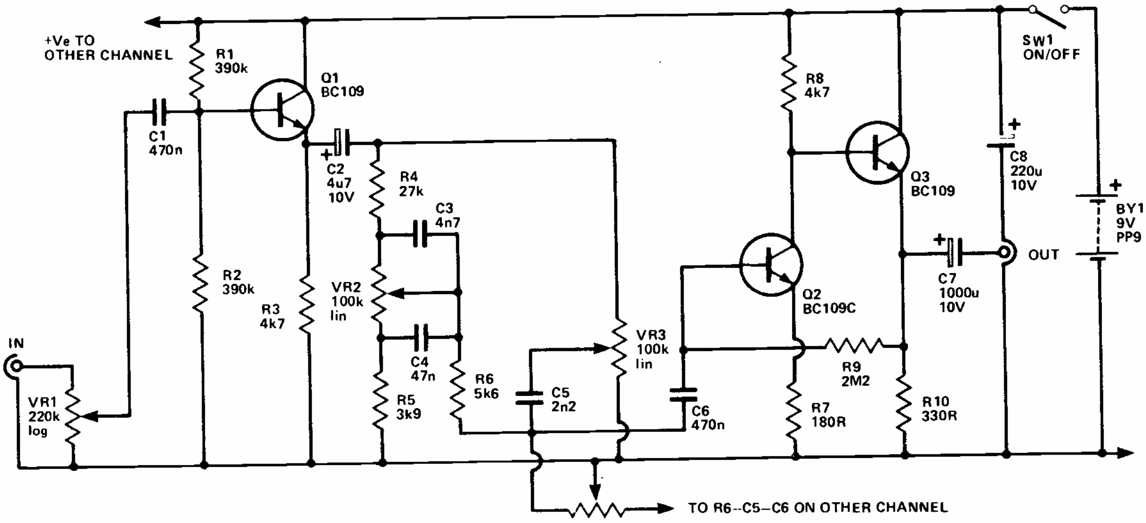
VOLTAGE GAIN = Rfb/Rin
NOTE: AMPLIFIER INVERTS SIGNAL

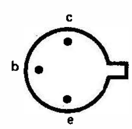
Headphone Amplifier

This simple stereo amplifier will drive a pair of stereo headphones, and can take its input from either a tuner or cassette deck.

The circuit shown here is for one channel, all the components being duplicated in the other channel except for S1, BY1, and VR4, which are obviously common to both channels. The two VR1s are a dual gang component, as are the two VR2s and the two VR3s.

The input signal is applied to volume control VR1, and from here it is coupled to a buffer stage based on Q1. This gives the unit a reasonably high input impedance of at least 100k. Its output feeds a conventional passive tone control circuit that can give bass lift or cut using VR2, and treble lift or cut using VR3. VR4 is used in the standard balance control arrangement. The output from the tone controls is coupled by C6 to a two stage direct coupled amplifier. This uses Q2 in the common emitter mode to give sufficient voltage gain for an output level of up to about 2Vrms from most sources. Q3 is an emitter follower buffer stage which matches the output from Q2 to the relatively low impedance of the headphones.

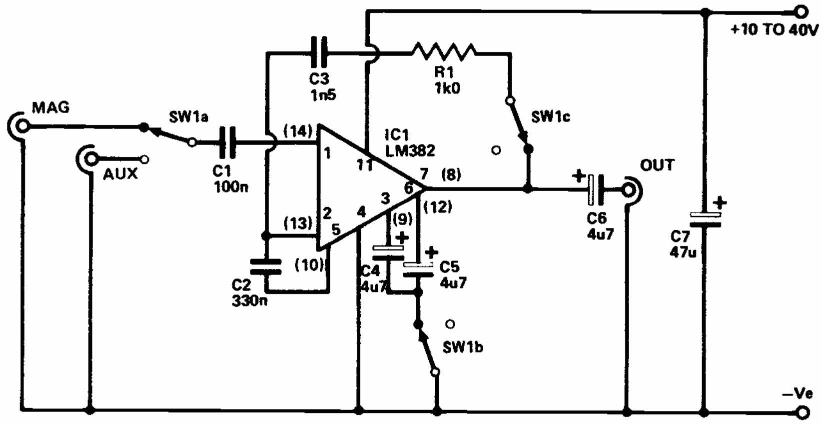




BC109 AND BC109C
BASE VIEW

Audio

Simple Preamplifier



This preamplifier has two inputs; one for a magnetic cartridge and the other is an "Aux" input for a tuner, tape deck, etc. Although the circuit is very simple, it uses an IC which is specifically designed for this application and provides low levels of noise and distortion. The unit is suitable for stereo operation and both the required amplifiers are contained within a single LM382 IC.

The circuit diagram is for one channel only, but apart from the IC pin connections the other channel is identical. The numbers in brackets show the pin connections for the other channel. The supply connections of IC1 are common to both channels.

The LM382 has an internal biasing circuit which sets the quiescent output voltage at approximately 6V and no discrete biasing components are required. C6 provides DC blocking at the output. When switched to the "Mag" mode, external feedback components are required to shape the frequency response characteristic of the amplifier in the required way. Bass cut and treble boost are applied to the signals transferred onto records so as to prevent excessive low frequency groove modulations and give an improved signal to noise ratio. The pre-amplifier must give corresponding bass boost and treble cut in order to give a flat overall frequency response. C2, C3 and R1 are the discrete feedback components and the LM382 itself contains some feedback resistors. C4 and C5 provide DC blocking for two shunt resistors in the feedback network.

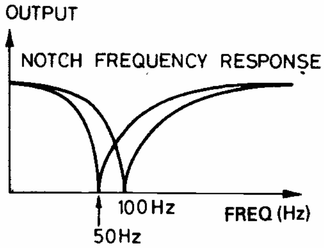
When switched to the "Aux" mode, most of the feedback components are not required and are switched out of circuit by SW1a and SW1c. C2 is left in circuit, but is superfluous. In this mode the circuit has a voltage gain of only about four and is really just operating as a buffer stage. SW1a connects the input of the amplifier to the appropriate input socket and C1 provides DC blocking at the input. Of course the input wiring must all be screened to prevent stray pick-up of mains hum, etc.

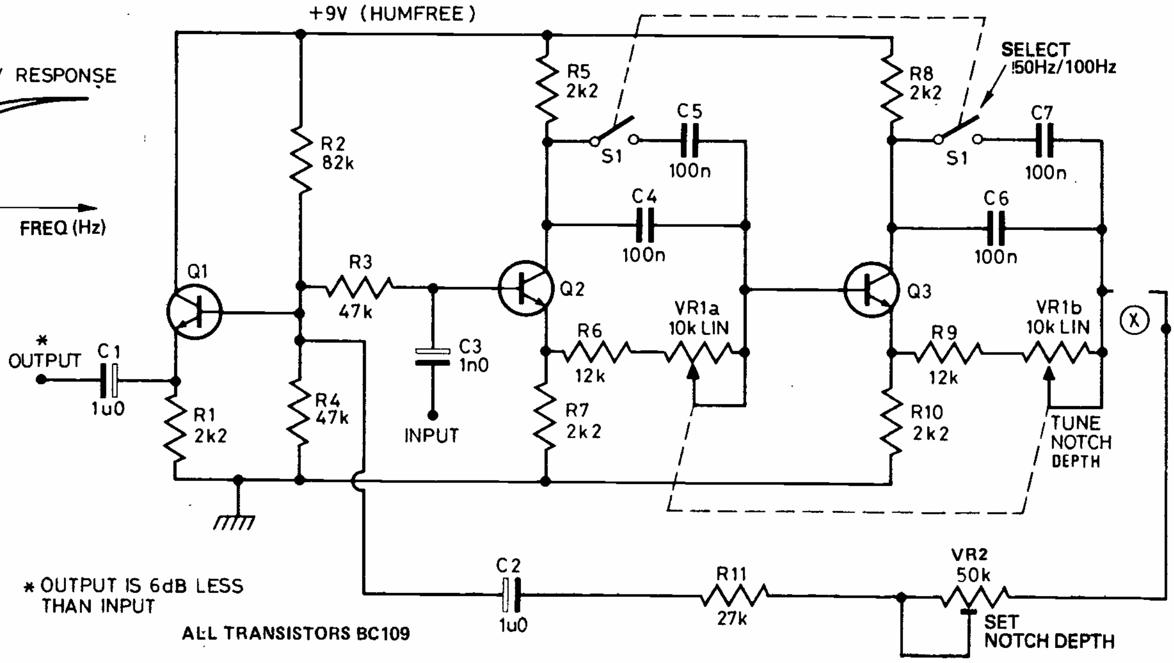
The current consumption of the circuit is about 12.5mA. Due to the high supply ripple rejection of the LM382, it is not necessary to have a highly smoothed and decoupled supply.

Hum Notch Filter

Magnetically induced hum is generally at a frequency of 50Hz, while ripple hum is mainly at 100Hz. This circuit can deal with either type with switch selection of 50Hz or 100Hz operation.

The filter consists of two transistor stages which each delay the signal by 90°. The total delay through Q2 and Q3 is therefore 180° at point X. The delay only affects the hum part of the signal and this delayed (or rather, phase shifted) hum is mixed back into the main signal via RV2. The hum in the main signal is cancelled by the 180° shifted hum, and the other (wanted) signal is virtually unaffected. RV1a, b is used to vary the centre of the notch frequency by plus or minus 40%. With the switch open circuit the frequency of operation is 100Hz.





* OUTPUT IS 6dB LESS THAN INPUT

ALL TRANSISTORS BC109

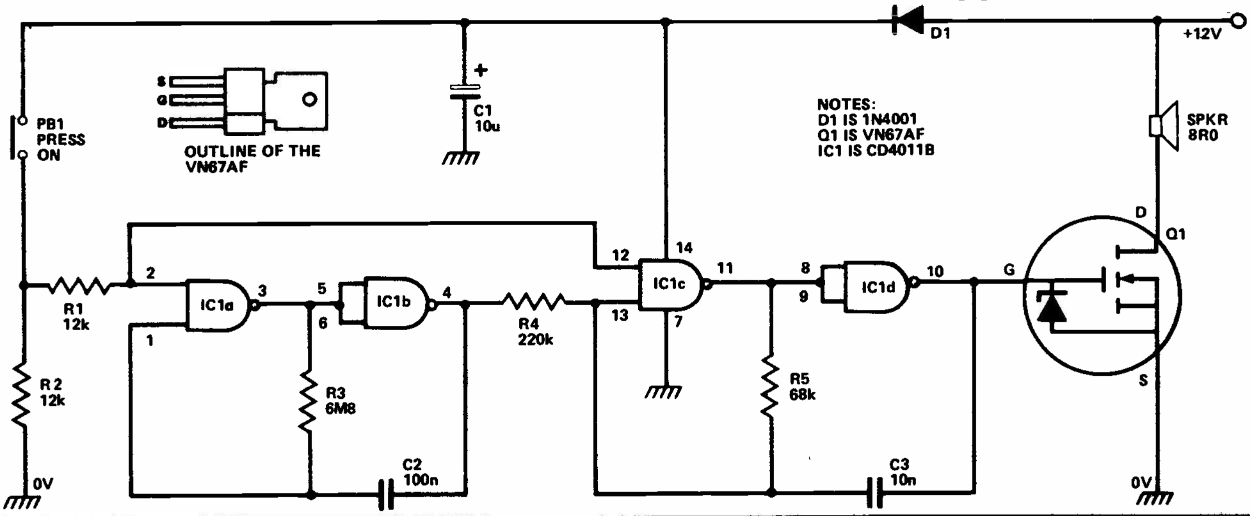
6 Watt Siren

This nifty little design uses the very latest advances in semiconductor technology to implement a very compact, inexpensive, yet exceptionally powerful alarm-sound generator unit that can easily be incorporated into an existing burglar alarm system or similar "security" device.

The alarm circuit incorporates a basic alarm-signal generator, followed by a power amplifier stage. The alarm-signal section of the unit is designed around an inexpensive CMOS integrated circuit that consumes virtually zero "standby" power. The power amplifier stage is a real state-of-the-art device, a low-cost VMOS power FET which also consumes virtually zero current when in "standby" mode. Consequently, the unit does not need a separate on/off switch and can be left permanently connected to a 12-volt battery supply.

IC1a and IC1b are wired as a slow astable multivibrator and IC1c-IC1d are wired as a fast astable. Both these astables are "gated" types, which can be turned on and off via PB1. The output of the ICa-IC1b slow astable is used to modulate the frequency of the IC1c-IC1d fast astable, and the output of the fast astable is fed to the external speaker via the Q1 VMOS power amplifier stage.

Normally, with PB1 open, both astables and Q1 are inoperative and the circuit consumes virtually zero standby current. D1 and C1 are used to ensure that the astable actions are not adversely influenced by voltage transients induced into the battery supply leads via the speaker. Note that the speaker used in the system must be an 8R0 type with a power rating greater than 6 watts.



Remote and Touch Volume Controls

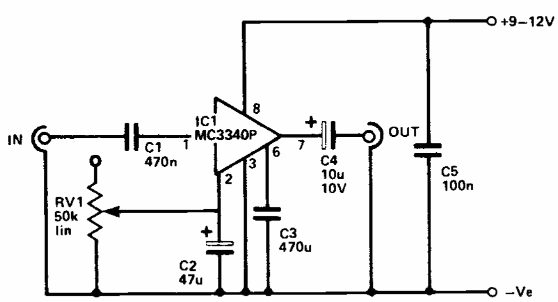
The MC3340P IC can be used as the basis of a remote volume control, as shown in the first circuit. RV1 controls the voltage gain of the MC3340P, which varies from typically 13dB at minimum resistance to about −80dB at maximum resistance. Since only a DC level is controlled by RV1, any AC pick-up in the connecting cable can be filtered out, which is the purpose of C2. C1 and C4 are input and output DC blocking capacitors respectively. C3 rolls off the RF response of the circuit to aid stability and prevent RF breakthrough.

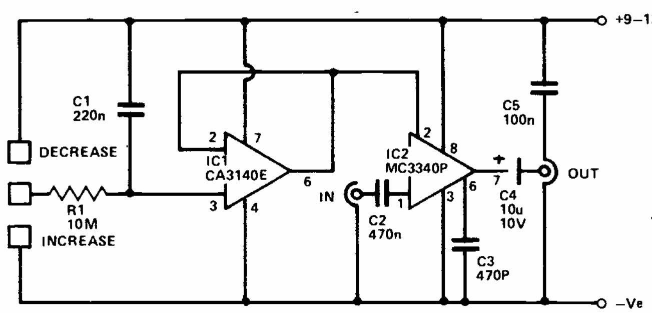
The MC3340P can be used as the basis of a novel touch operated volume control, as shown in the second circuit. The device is controlled by a voltage rather than a resistance and gives the same attenuation range as the previous circuit.

The control voltage is obtained from a charged capacitor (C1) via an op-amp unity gain buffer stage utilising IC1. IC1 has a CMOS input stage which produces a typical input resistance of 1.5 *million* Meg ohms. This ensures the charge on C1 is not significantly affected by the circuit; once set, it remains virtually unaltered for a long time.

The charge on C1 is set by the operator who, touching the lower two contacts, can charge C1 via R1 and his or her skin resistance. This decreases the control voltage fed to IC2, and increases the volume. Touching the upper two contacts causes C1 to gradually discharge; increasing the control voltage and decreasing the volume. When the unit is switched off, C1 gradually discharges. At switch-on it is necessary to bring the volume up to the required level, rather like using an ordinary combined on/off switch and volume control.

Both circuits will handle input levels of up to 500mVrms, with a THD figure of about 0.6% at high volume settings, rising to about 2% or so at low settings.





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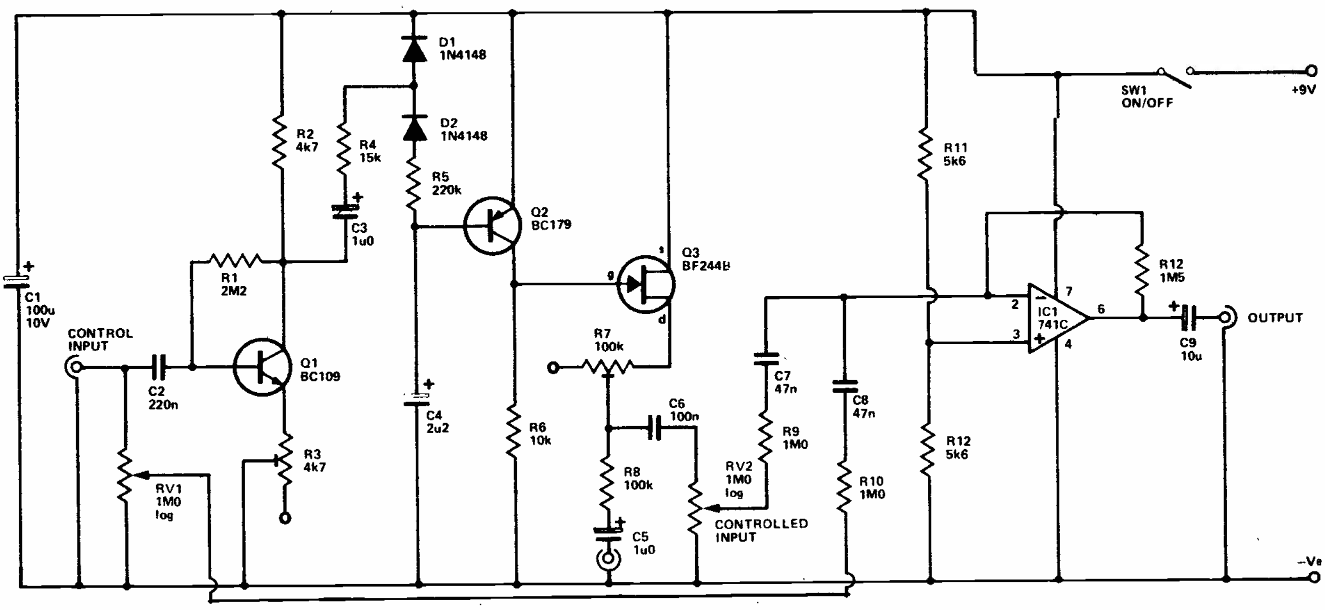
Voice Operated Fader

The purpose of a voice operated fader is to mix two signals and to automatically fade out one of these to some predetermined level when a signal is present in the other channel. Normally the control signal is a commentary of some kind and the controlled signal is music. With no control input signal, the controlled signal is fed via C5 and R8 to one input of a conventional two input mixer which is based on IC1.

Some of the control signal is applied to the other input of the mixer, and the rest is fed to a common emitter amplifier based on Q1. The output of this is fed to a rectifier and smoothing circuit comprised of C3, R4, D1, D2, R5 and C4. Normally Q2 is cut off, and R6 strongly reverse biases JFET Q3. Q3 therefore has a very high drain to source resistance (typically about 1,000 megohms), and has no significant effect on the circuit. However, in the presence of a suitable control input signal, the negative bias produced by the smoothing and rectifier circuit switches Q2 hard on, and eliminates the reverse bias on Q3. The drain to source resistance of Q3 then falls to about one hundred ohms or so and causes greatly increased losses through R8. Thus the required automatic fade out action is obtained. R7 can be adjusted to give any level of fade out from about -6dB. to about -60dB.

R3 controls the voltage gain provided by Q1, and in practice is adjusted for the lowest gain (highest resistance) that gives reliable operation of the unit. Note that this input is not intended to operate direct from a microphone, and a suitable pre-amplifier must be added here.

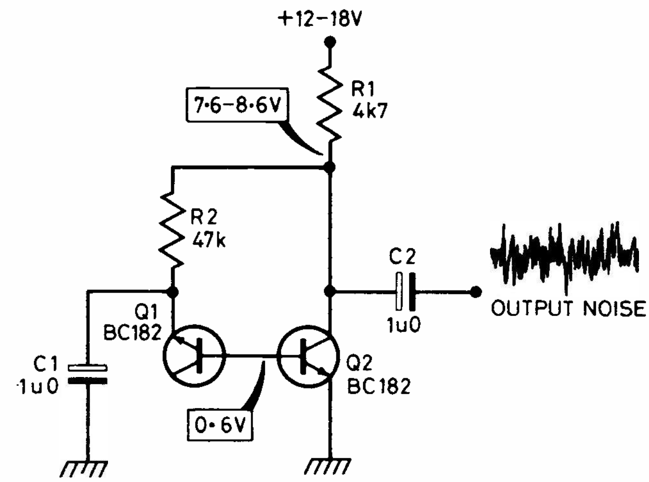


Noise Source

The sort of noise we mean is not the type neighbours complain about. This circuit produces an electrical signal which, when suitably amplified, sounds like a hiss of escaping gas. This can be used for sound effect production – it can form the basis of sound effects ranging from surf hitting a beach to a steam train – or for testing loudspeakers.

The operation of the circuit depends on the reverse-bias breakdown of Q1. This occurs when the voltage across the emitter and base of the transistor reaches 7 to 8 volts. At this voltage the transistor produces noise. The rest of the circuit is dedicated to keeping the current through Q1 to just the right level (too little – no noise; too much – dead transistor!) and to amplifying the result.

The bias for Q2 comes through Q1; if Q1 passes a lot of current,

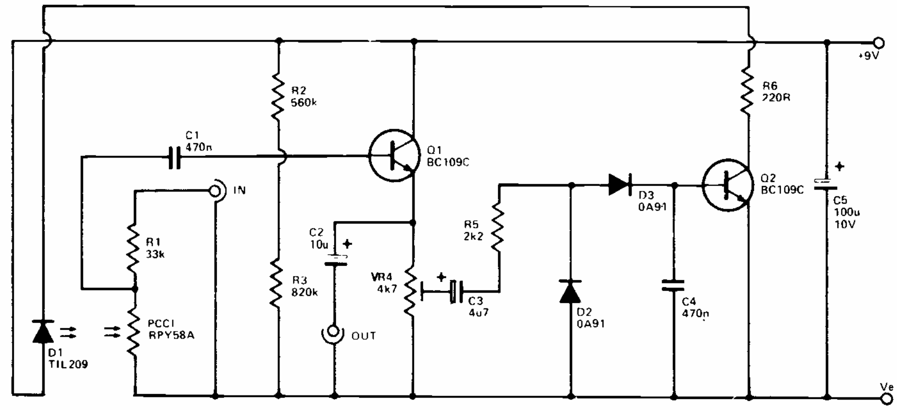


Q2 will turn on more and the voltage at the bottom of R1 will drop. This will cause the voltage across Q1 to drop and the current through it will decrease. In this way the current is kept to a reasonable level. C1 provides a path to earth for the high frequency noise which appears on the collector of Q2. This ensures that the fast changes in the current through Q1 (and this is, after all, what we are after) are not 'adjusted' in the same way and lost.

Audio Limiter

When making tape recordings, especially of "live" performances, it can be very difficult to set the correct recording level. This can easily lead to an excessive recording level and consequent distortion occurring unless the recording level control is kept well backed off. The price one then has to pay is a low recording level and subsequent low signal to noise ratio. The normal way of overcoming this problem is to use an audio limiter circuit ahead of the tape deck. This device normally passes the signal straight through to the recorder, but if the input exceeds a preset threshold level it attenuates the signal so that the output level is not sufficient to overload the recorder.

In this circuit the input signal is applied to an attenuator which is formed by R1 and PCC1. Normally PCC1 is in total darkness and exhibits a very high resistance (typically a few megohms) causing minimal losses through the attenuator. This stage feeds into the high input impedance of the emitter follower buffer stage formed by Q1 and its associated components. and this ensures little loss of signal level.



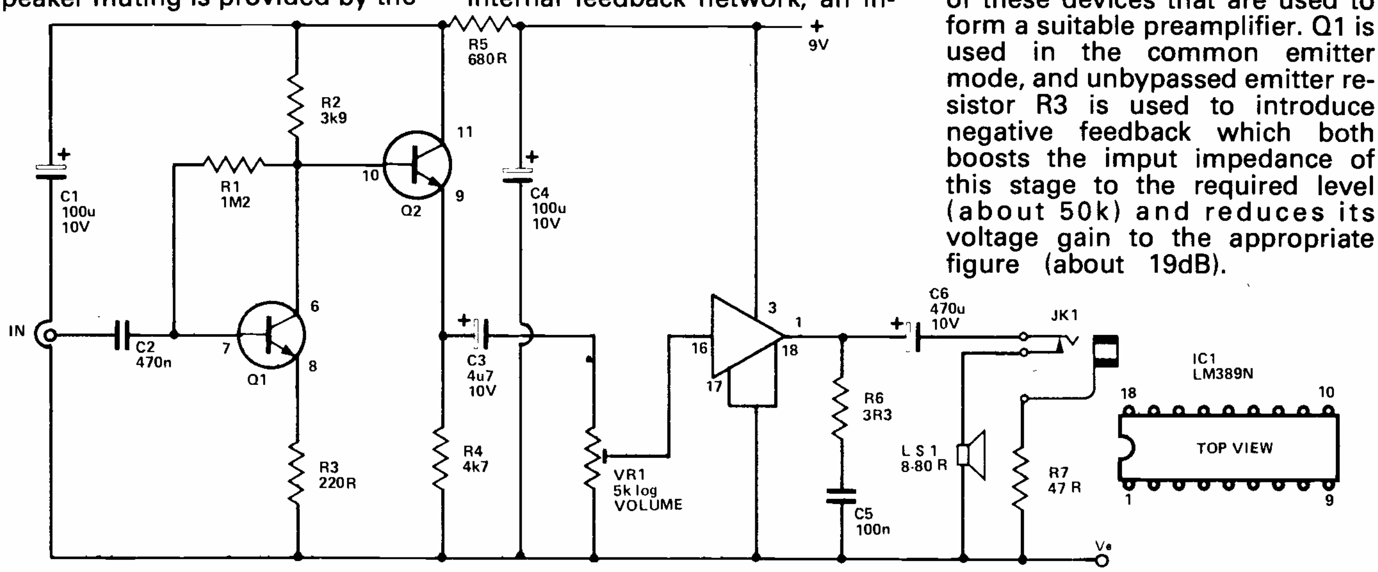
Some of the output signal is fed from the slider of VR4 to a rectifier and smoothing circuit which is composed of D2, D3, and C4. If the input signal is sufficiently strong, the positive bias produced by the circuit will be adequate to switch on Q2 and light emitting diode D1 which is connected in its collector circuit. The light output from D1 is aimed at the sensitive surface of PCC1 (the surface to which the leadout wires do not connect) and this causes a large reduction in the resistance of PCC1. The larger the input signal is made, the more strongly D1 glows, and the greater the recuction in gain. This has the effect of preventing the output level from rising far above the level at which D1 begins to initially switch on. On the prototype this threshold level is at about 230mV and increasing the input level to 4 volts rms causes the output to rise to only about 320mV. Higher threshold levels can be obtained with VR4's slider adjusted to the appropriate point. The attack and decay times of the circuit are both quite short so that the unit quickly responds to changes in signal level and is not normally conspicuous in operation.

Construction of the unit should be quite straight forward, but the unit must of course be housed in a light proof box so that PCC1 is shielded from the ambient lighting.

Guitar Practice Amplifier

The circuit is based on an LM389N audio amplifier IC which contains a small class B power amplifier. The output of the amplifier is coupled to the speaker via DC blocking capacitor C6 and the break contact of headphone socket JK1. Automatic loudspeaker muting is provided by the latter when the headphones are connected. The output voltage swing is considerably in excess of that required by most phones, and so a degree of attenuation is provided by series resistor R7. R6 and C5 are a Zobel network which aid the stability of the circuit.

As the voltage gain of the amplifier is preset at about 26dB by an internal feedback network, an input level of over 100mVrms is required to produce full output. Some form of preamplifier is therefore required. The LM389N has the unusual feature of containing three NPN transistors in addition to the power amplifier circuit. The terminals of these transistors are connected to individual pinouts of the device, and it is two of these devices that are used to form a suitable preamplifier. Q1 is used in the common emitter mode, and unbypassed emitter resistor R3 is used to introduce negative feedback which both boosts the imput impedance of this stage to the required level (about 50k) and reduces its voltage gain to the appropriate figure (about 19dB).



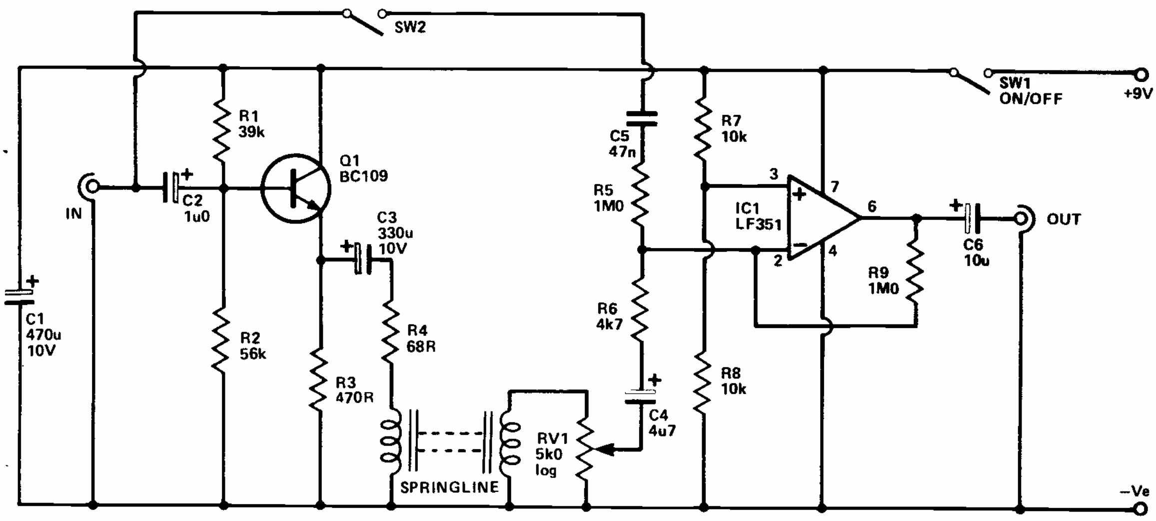
Reverberation Unit

This unit simulates the long reverbation time of a large hall (usually around 2s or so) and can be employed as a musical effects unit or to improve certain types of home-recording. Reverberation is caused by sounds being reflected around the interior of a room and in the case of a large hall the sounds are usually reflected many times before losing sufficient energy to render them inaudible. This, coupled with the fairly long distances covered by the sound waves between reflections, gives the long reverbation time and reverberant sound of a large hall.

There are several ways of simulating reverberation, but the simplest and most commonly used is probably the springline system. A springline consists of two transducers linked by one or two long springs. If a signal is fed into one transducer it produces a corresponding audio signal which is transmitted down the spring to the second transducer. Here it is reconverted into an electrical signal again. However, the sound signal travels down the spring relatively slowly, and the signal is reflected backwards and forwards along the spring many times before it decays to an insignificant level. Thus, the output from the second transducer is a good simulation of natural reverbation.

In this circuit the input signal is fed to the low impedance input transducer of a short springline via an emitter follower which gives a reasonably high input impedance of about 10kR or so. This uses Q1 in a conventional configuration. The output of the springline unit is fed to one input of a mixer circuit. This is based on IC1 and again uses a conventional and well known arrangement. There are substantial losses through the springline and so the mixer is designed to boost the output of the springline by over 46dB (200 times). The other input of the mixer is fed with the input signal, but the high value of R5 gives only about unity voltage gain at this input, so that the main signal does not overwhelm the reverberation signal.

RV1 enables the amount of reverberation signal mixed into the main signal to be controlled. It can be reduced right down to zero by fully backing off RV1. SW2 can be used to cut out the main signal so that only the reverberation signal appears at the output, if desired. The only other control is on/off switch SW1. The current consumption of the unit is approximately 10mA.

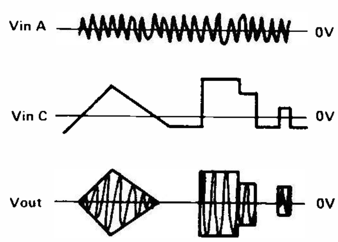


CA3080 VCA

A simple voltage controlled amplifier can be made using a CA3080, an operational transconductance amplifier made by RCA. This is basically an op-amp with an extra input at pin 5. A current I is injected into this input and controls the gain of the device linearly. Thus by inserting an audio signal (±10mV) between pin 2 and 3 and by controlling the current on pin 5, the level of the signal output (pin 6) is controlled.

In effect, the audio signal is multiplied by the current I. The output of the CA3080 is a current output and so a resistive load (R5) is needed; R5 in fact becomes the output impedance of the circuit.

The current controlling the CA3080 is generated with a

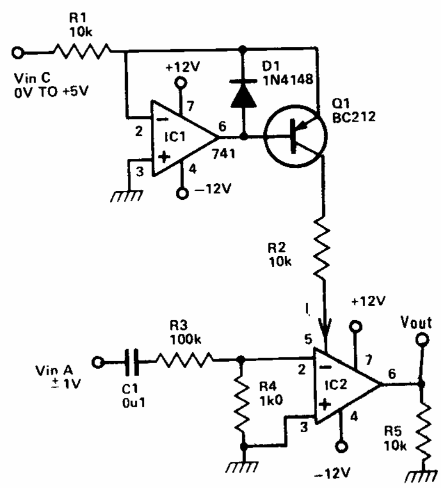


NOTE: NO OUTPUT WHEN Vin C IS NEGATIVE

voltage to current converter IC1, Q1, R1. This circuit linearly converts VinC into a current (I) where:

$$I = \frac{VinC}{R1}$$

When VinC=OV, I is O so the CA3080 is turned off. When VinC is positive, I is generated and so the VCA is turned on. When VinC goes negative, I is off, so no output is produced from the VCA.



The VCA finds many uses in the automatic control of signal levels and in generating envelope contours in electronic musical equipment.

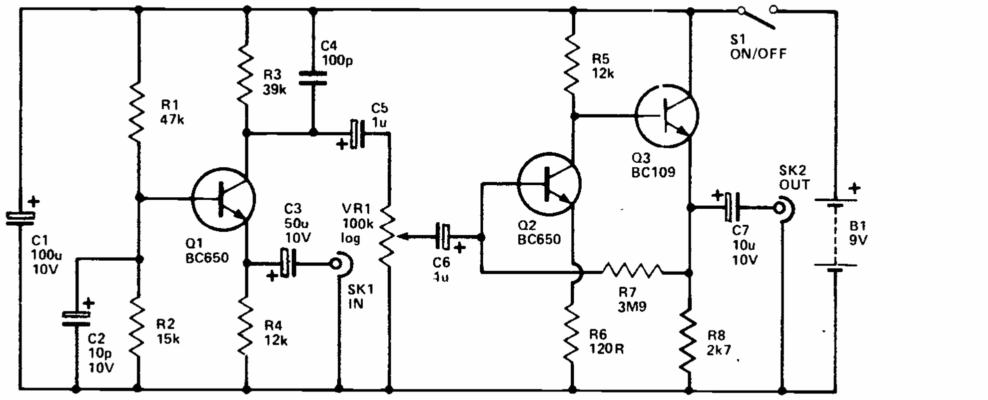
Music and Disco

Two Microphone Preamplifiers

The first preamplifier circuit shown here is for use with low impedance (200 or 600 ohms) dynamic microphones. Most electret type microphones have an integral source follower JFET buffer stage which gives a low output impedance, and should also work well with this circuit. Low impedance microphones have an extremely small output voltage which in normal use is unlikely to ever exceed 1mVrms. This means that the preamplifier must have a high voltage gain and a very low noise input stage if a high signal to noise ratio is to be achieved. This circuit requires only about 200uV rms at the input for an output level of 1V rms, and the unweighted signal to noise ratio (input open circuit) is a little under −66 dB.

A common base input stage is used, based on Q1 which is an ultra low noise, high gain device. Although the common base configuration is not often encountered in audio circuits it is ideal for this application as it gives the required low imput impedance and high voltage gain. In order to obtain a really low noise level Q1 is operated at a collector current of about 180uA. C4 is an RF filter capacitor, and reduces the risk of radio signals picked up by the input wiring breaking through to the output. With high gain circuits such as this RF breakthrough is not an uncommon problem, and ideally the unit should be built into a metal case so that the circuit is screened from RF signal sources.

The output from Q1 is fed by way of gain control RV1 and coupling capacitors C5 and C6 to a two stage amplifier circuit. Q2 is a common emitter amplifier and provides additional voltage gain. Q3 is used as an emitter follower output stage and gives the circuit

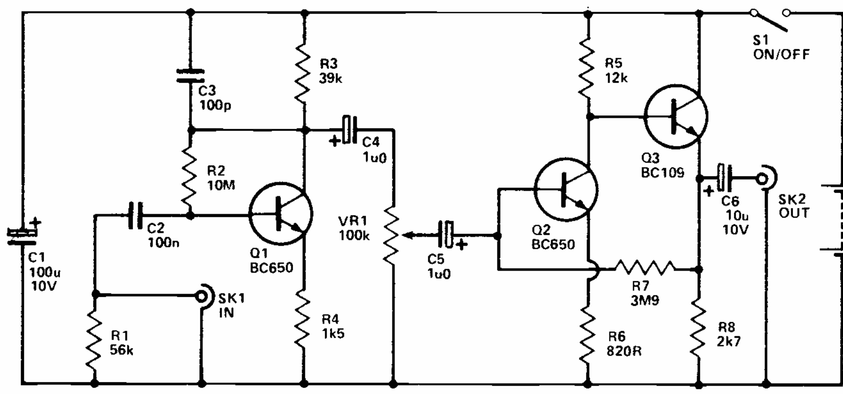


a low output impedance. Power is obtained from a 9 volt battery supply and the current consumption is about 3 mA.

Many dynamic and electret microphones have an integral step up transformer which gives increased output voltage, but the available output current is of course decreased and so a preamplifier having a higher input impedance (usually 50k) is required.

The high impedance microphone preamplifier shown here requires an input level of approximately 5mVrms for an output of 1Vrms, and the unweighted signal to noise ratio (input short circuited) is well over −70 dB, reference this output level. Apart from the input stage the circuit is virtually identical to the previous design, the only difference being that the emitter resistor for Q2 has been increased in value. This has been done because the circuit only needs to have a moderate amount of voltage gain, and the increased negative feedback produced by raising the value of the resistor gives the necessary reduction in gain.

The input stage is again run at a fairly low collector current (about 250uA.) in order to give a low noise level. Only a moderate voltage gain and medium input impedance are required from this stage, and so Q1 is employed in the common emitter mode with non-bypassed emitter resistor R4 being used to introduce negative feedback, to give increased input impedance and reduced voltage gain. The input impedance to Q1 is actually somewhat higher than is required, and so R1 is used to shunt the input and reduce the impedance to approximately the required figure of 50k.

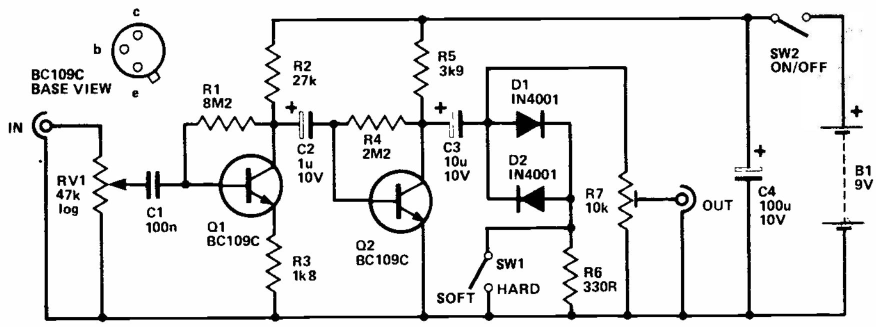


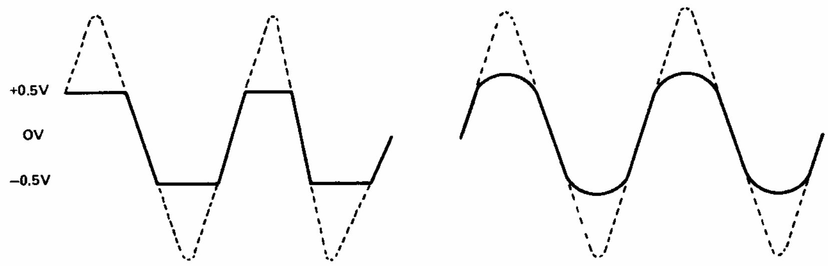
Clipping Amplifier

Probably the main use for clipping amplifiers these days is in musical "fuzz" effect. This circuit uses two common emitter amplifiers based on Q1 and Q2 to drive a simple clipping circuit using D1 and D2. RV1 is the input attenuator and if this is adjusted for an output level of less than about 1 volt peak to peak at Q2 collector, neither D1 or D2 will be sufficiently forward biased to conduct significantly. These components then have no real effect on the circuit, which in consequence operates as an ordinary amplifier. Assuming SW1 is closed, and if RV1 is adjusted for a signal level of more than 1 volt peak to peak at Q2 collector, during positive output excursions when the signal amplitude is greater than 0V5, D1 will conduct and act like a low voltage zener, preventing the signal from exceeding 0V5 in amplitude. Similarly, on negative output excursions D2 will limit the signal level to no more than −0V5. This causes the signal to be severely distorted by the clipping action as shown in (a), the distortion products giving the desired "fuzz" effect.

A circuit of this type can be used

to produce a form of sustain effect when employed with a guitar. Here RV1 is adjusted so that clipping occurs even when the signal from the guitar has decayed considerably. This results in the output signal remaining at a virtually constant 1 volt peak to peak level for the duration of each note, whereas a guitar signal normally hits a high initial peak and then rapidly decays. In this application the hard clipping produced by the unit will produce the fuzz distortion products whether they are required or not. This problem can be alleviated to some degree by switching SW1 to the "soft" position. R6 is then connected in series with D1 and D2, and this gives the smoother clipping action shown in (b) due to the voltage developed across R6 when D1 and D2 pass a current. This greatly reduces the high frequency distortion products, the most noticeable and objectionable ones.





The circuit has an input impedance of about 47kR and needs an input of less than 1mV rms to produce clipping. If the full 1 volt peak to peak output is not required, R7 can be used to attenuate the signal to the required level.

Speech Processor

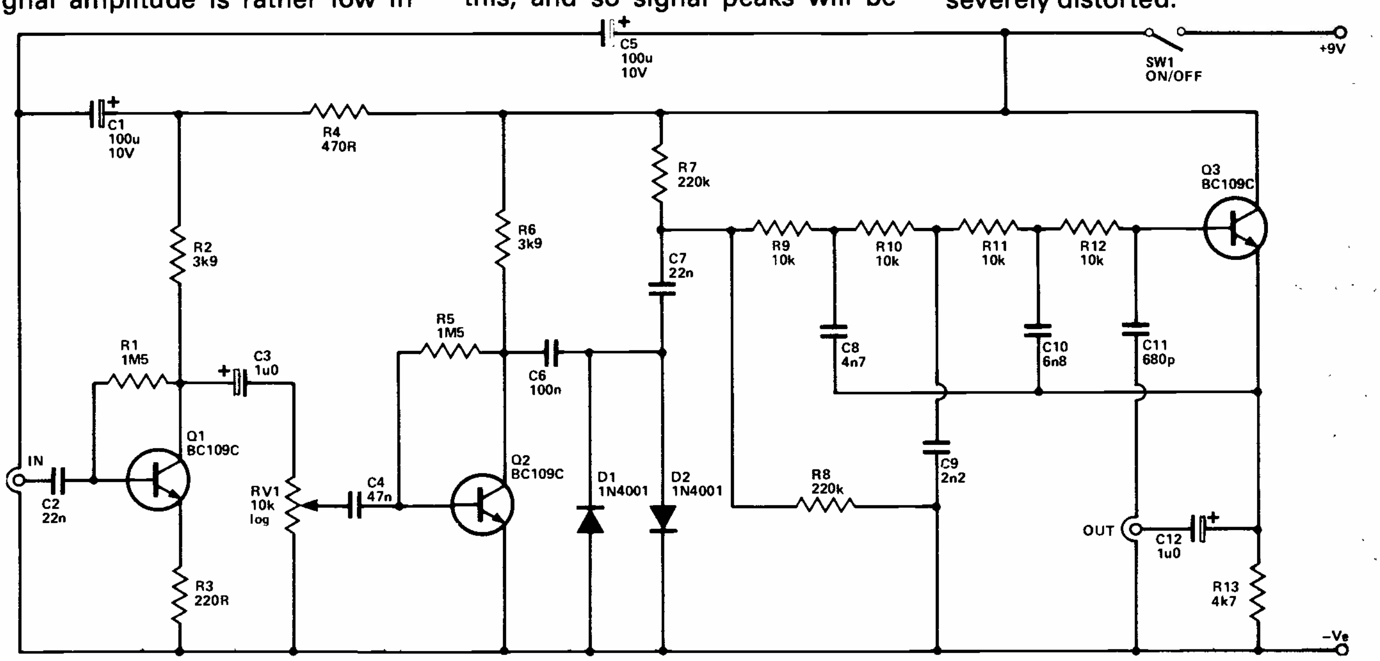
Speech processors can be used in public address systems, disco equipment, etc. Bass frequencies do not aid the intelligibility of speech, and treble frequencies are of little help either. The signal can therefore be made more effective by attenuating the higher frequencies (above about 3kHz) and the lower frequencies (below about 250Hz), and then boosting the signal to compensate for the reduction in its amplitude.

A speech signal has a rather spikey waveform, and the average signal amplitude is rather low in comparison to the peak level; the signal is given considerably more "punch" if the peak levels are compressed.

In this circuit the signal from the (50k impedance) microphone is amplified by a two stage pre-amplifier using Q1 and Q2 in the common emitter mode. RV1 is microphone gain control. The output from the preamplifier is coupled by C6 to a simple clipping circuit which uses D1 and D2 to clip the signal at about ± 0V5. In use the output from the preamplifier will be somewhat higher than this, and so signal peaks will be severely reduced by the clipping circuit, but signals below ± 0V5 will not be affected.

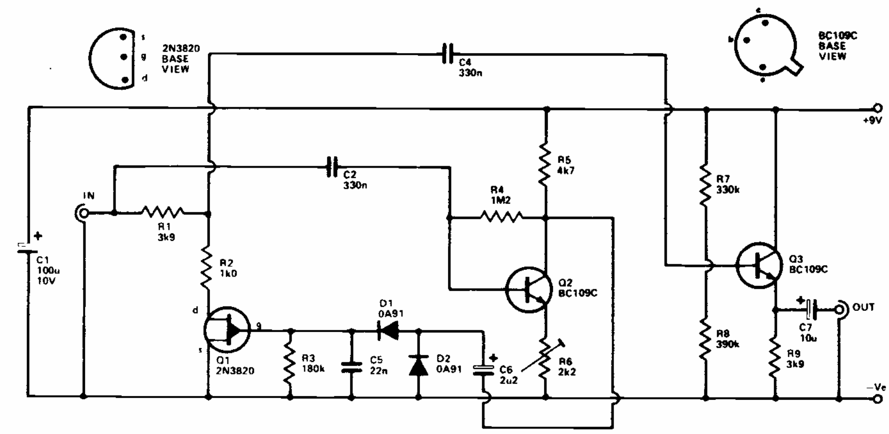
This gives the required compression but also generates high frequency distortion components. These are attenuated by an active low pass filter based on Q3 and having a nominal 18dB per octave roll off. The low frequency attenuation is obtained by using low value coupling capacitors.

RV1 is adjusted to give the desired amount of compression, and should not be advanced too far as this would cause the output to be severely distorted.



Music and Disco

Noise Gate



The purpose of a noise gate is simply to reduce the gain of the equipment when there is only a low signal level. Thus, during brief pauses that occur during normal speech the background noise is attenuated but, while the speech signal is present, the gain of the equipment is returned to its normal level. The true signal to noise ratio is not actually improved at all, but the monotony of the continuous high noise level is removed, making the signal easier to listen to and aiding intelligibility.

Q1 is an N channel JFET which is used here as a voltage controlled resistance. Under quiescent conditions Q1 gate is biased to the negative supply rail by R3, and this switches the device hard on so that it exhibits a low resistance of about 100 ohms. The input signal is applied to an attenuator which has R1 as the series element, and R2 plus the drain to source resistance of Q1 as its shunt element. With Q1 switched on there is a loss of about 11dB through the attenuator.

Some of the input signal is amplified by a common emitter amplifier based on Q2, and then fed to a rectifying and smoothing circuit which is comprised of D1, D2, and C5. If the input signal is of sufficient strength this produces a strong positive bias to switch off Q1, causing its drain to source resistance to increase to many megohms. There is then very little attenuation of the signal. Thus, low signal levels (noise only) are reduced while high signal levels (noise and wanted signal) are allowed to pass virtually unaffected. The attenuator must feed into a fairly high impedance or the action of the circuit will be impeded. An emitter follower buffer stage based on Q3 is therefore interposed between the attenuator and the output. The time constant of the circuit has been made quite short so that there is a quick response to changes in signal level.

The circuit will operate with an input level of between about 50mV and 2Vrms. R6 is adjusted for the lowest resistance that does not cause the background noise to activate the unit.

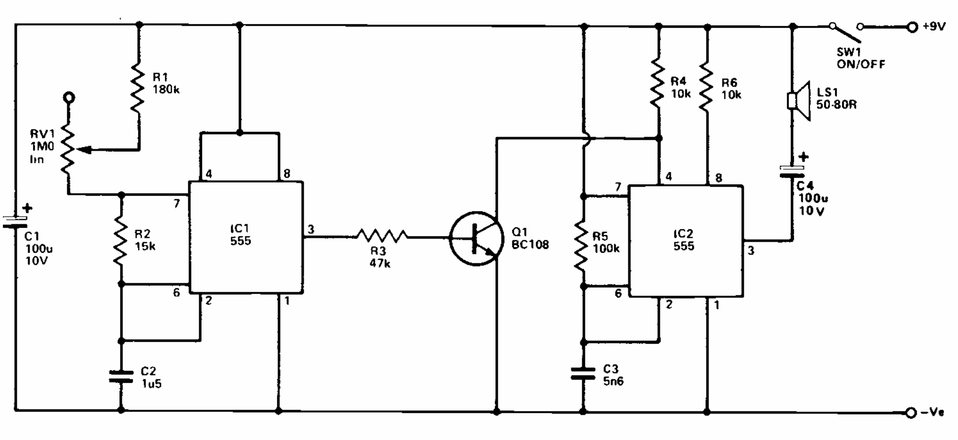
Metronome (1)

There have been a great number of electronic metronome designs published over the last few years, the majority designed to simulate the sound of a mechanical metronome (Maelzel's metronome). The usual method of achieving this consists of feeding brief pulses to a loudspeaker at the appropriate beat rate, giving the required "clicking" sounds. One drawback of this system is that a normal miniature loudspeaker can only give limited volume from such a signal.

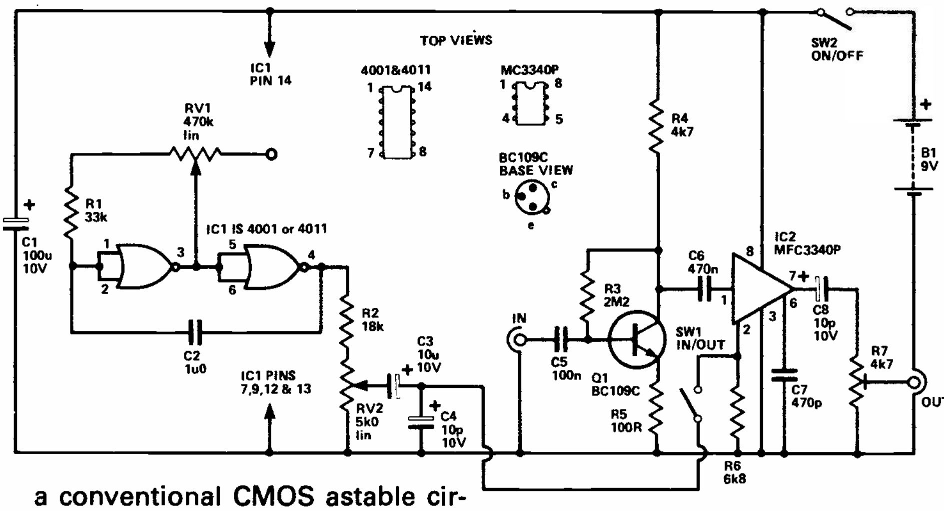
One way of producing a more effective metronome that is not easily masked by the sound of music, is to feed the speaker with a pulsed tone, giving a much more noticable and penetrating sound. The sound produced is somewhat different to that of a conventional metronome, but is nevertheless perfectly acceptable in practice.

The circuit shown here is for a metronome of this type. The audio tone is produced using a 555 astable circuit based on IC2 and its associated components. The timing components (R4, R5, and C3) give a roughly squarewave output at a frequency of about 1k24Hz. This signal is fed to the loudspeaker via DC blocking capacitor C4.

IC1 is used in another astable circuit which is used to pulse the tone generator at the required beat rate. The operating frequency of this oscillator can be varied from about 290 pulses per minute with RV1 at minimum, down to about 48 per minute when it is set at maximum. The timing component values have been chosen to give very brief negative output pulses; during these, Q1 becomes cut off. Normally Q1 holds pin 4 of IC2 at a fraction of a volt and prevents the tone generator from oscillating, but this muting is removed when Q1 becomes cut off, and a brief tone burst is fed to the speaker.



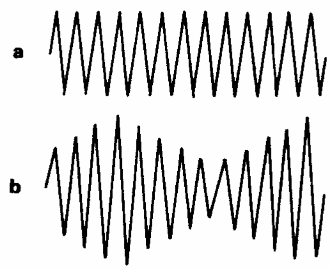
Tremolo Unit



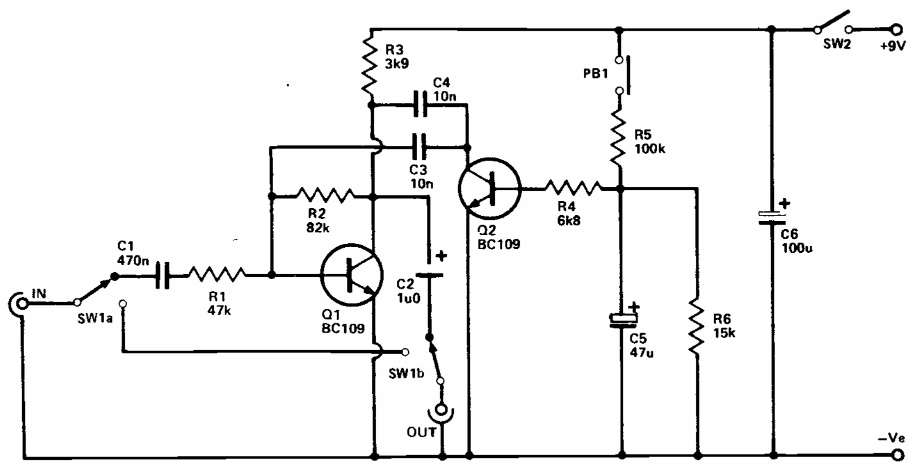
This is one of the most popular types of special effect unit for use with guitars. The operation is to amplitude modulate the input signal with a low frequency signal. Thus a constant input as in (a) would emerge from the tremolo unit varying in amplitude at a low frequency as in (b).

In this circuit the input signal is taken to the input of an electronic attenuator (based on IC2) via common emitter amplifier Q1. R6 sets the gain of the attenuator (with zero modulating voltage) at about unity, but the amplification provided by Q1 gives an output level of a few hundred millivolts. This can either feed a high level amplifier input or R7 can be adjusted to attenuate the output to a level which is suitable to drive the ordinary guitar input. It is necessary to have the amplification stage ahead of IC2 so that this part of the circuit is handling a fairly high signal level and gives a good signal to noise ratio.

The gain of IC2 can be varied by applying a control voltage to pin 2. This control signal is generated by a conventional CMOS astable circuit which uses two of the gates contained in IC1. The operating frequency of the astable can be varied from about 1 to 10 Hz by means of frequency control RV1. A squarewave signal is produced by the astable, and this must be filtered to remove the high frequency components in order to give a smooth and pleasant tremolo effect. This filtering is given by R2 and C4. RV2 controls the amplitude of the modulating signal and acts as the tremolo depth control. SW1 can be used to disconnect the modulation when the tremolo effect is not required.



Auto Waa



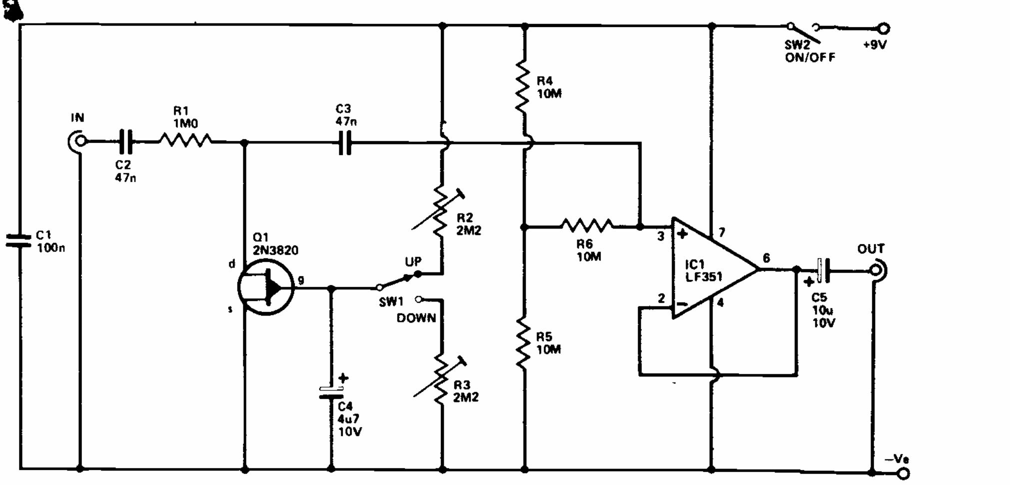
An unusual feature of this circuit is that the Waa-Waa effect is obtained by operating a foot-switch, rather than the more usual method of operating a potentiometer via a pedal mechanism.

The circuit uses a quite conventional arrangement based on common emitter amplifier, Q1. Frequency selective negative feedback is provided by C3, 4. These provide little feedback at a certain frequency. A peak in the response of the amplifier is produced at this frequency, as the lack of feedback enables virtually the full voltage gain of Q1 to be realised. The actual frequency at which the peak is produced can be controlled by means of a resistance between the juction of C3, 4 and the negative supply rail. With a high resistance here the peak is produced at a high frequency. By varying the control resistance the peak can be swept up and down the audio frequency spectrum, producing the familiar Waa-Waa effect.

The control resistance is formed by the collector to emitter impedance of Q2. Under quiescent conditions Q2 is switched off and the peak is at such a low frequency that it is effectively non-existent. If PB1 is operated, C5 charges up via R5 and, as the voltage across C5 increases, Q2 is biased harder into conduction by the base current it receives through R4. This causes the peak to be swept up through the audio band until C5 becomes fully charged. If PB1 is then released, C5 gradually discharges through R4, Q2 and R6, causing the bias on Q2 to decrease and the peak to be swept down the audio spectrum. Thus the required effect is produced by closing and opening PB1. The Waa-Waa frequency is partially controlled by the frequency at which PB1 is operated, but C5 restricts the range of frequencies that can be obtained in practice. However, the value of C5 can be altered to suit individual requirements, or several switched components of different values could be used.

SW1 enables the Waa-Waa circuit to be quickly and easily bypassed. R1 is needed to reduce the gain of the unit which would otherwise be excessive. Current consumption is about 2mA.

Automatic Fader



This unit can be used for discos, slide shows, film shows, etc. At the flick of a switch it can be used to automatically fade a signal in or out, without introducing any clicks or other background sounds.

In this circuit P channel JFET transistor Q1 is used as a voltage controlled resistance. Its drain to source resistance forms a voltage controlled attenuator in conjunction with R1. The input signal is applied to this by way of DC blocking capacitor C2. When power is initially applied to the circuit, C4 will be uncharged and the gate to source voltage of Q1 is therefore zero. This gives Q1 a low drain to source resistance of only 100 ohms, causing high losses through the relatively high resistance of R1. The output from the attenuator is only one 10,000th of the input level (- 80dB) and the signal is thus effectively cut off. As C4 begins to charge via R2, Q1 becomes increasingly reverse biased, causing its drain to source resistance to increase. As this resistance increases, the losses through R1 decrease, causing the signal to "fade" in. Eventually the drain to source resistance of Q1 reaches its maximum value of about 1,000 megohms, which gives no significant losses through R1.

Switching SW1 to the "down" position gradually discharges C4 through R3, returning the bias on Q1 to its original state and fading out the signal.

IC1 is used as a high impedance buffer stage that ensures little loading is placed on the attenuator.

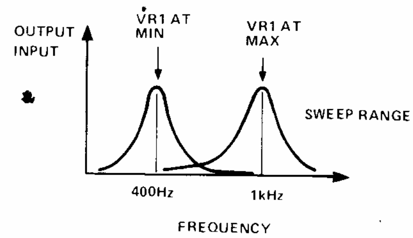
The fade up and fade down times are controlled by the settings of R2 and R3 respectively, and can be adjusted from a fraction of a second to about 8 seconds or so. If the full fade out level of about - 80dB is not required, a 1M preset can be inserted between Q1 drain and the junction of R1-C3; the fade out level can then be adjusted from about 6dB to the full 80dB. The current consumption of the unit is only 2mA.

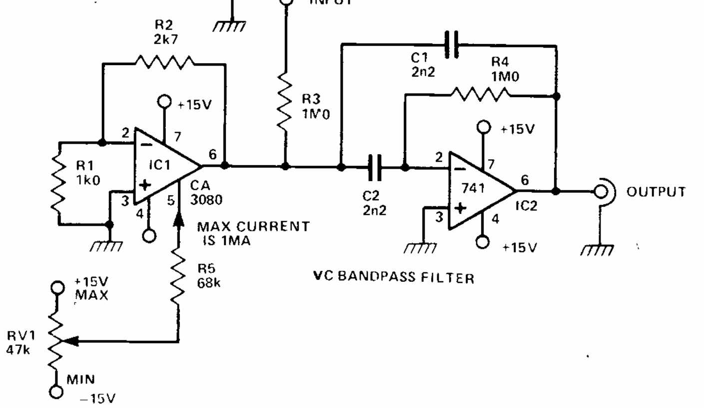
Voltage Controlled Filter

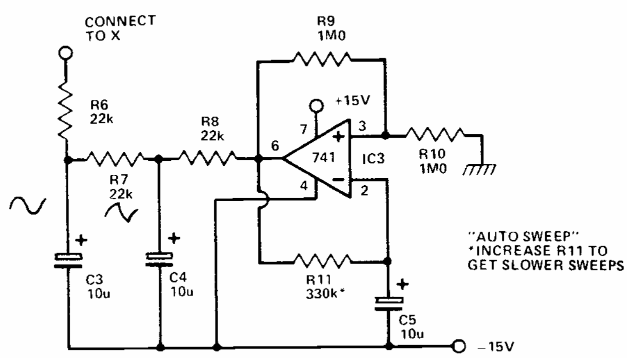
The CA3080 is an operational transconductance amplifier; what this means is that it is an amplifier whose gain can be controlled. Thus the CA3080 finds lots of uses in circuits where something has to be varied electronically, one such example being a voltage controlled filter.

A bandpass filter is constructed using IC2, R4, C1, C2. This is a multiple feedback bandpass filter and normally there is another resistor which is connected from ground to the junction of the two capacitors. By varying this 'other' resistor, the resonant frequency of the filter can be changed. The CA3080 and R1, R2 is this 'other' resistor. By varying the current into pin 5 of IC1 it is possible to control the gain of the device. R1, R2 provide negative feedback around the IC and this turns the network into a current controlled resistor. By increasing the current, the effective resistance is reduced, which in turn alters the resonant frequency of the filter.

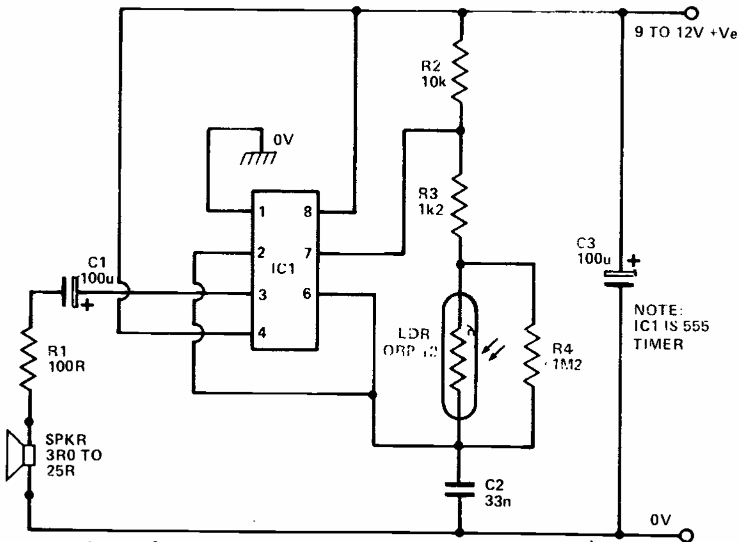
It is possible to provide a varying control current using the circuit involving IC3. This is a low frequency squarewave oscillator. The oscillation frequency is determined by R11,C5. The squarewave is heavily filtered by R6,R7,R8 and C3,C4, to produce a smoothly modulating current drive to IC1. This causes the centre frequency of the filter to be swept up and down. The autosweep can be used as an effect for electric guitar or for electronic music processing.







Opto-Tone



This circuit is a variable-frequency high-power oscillator that alters it's tone in response to the light intensity falling on the face of an LDR. The output tone is fed directly to a loudspeaker, and is light-variable over a three decade range. The tone is high at high levels of illumination, and low at low levels of illumination.

The circuit can be used as a simple musical instrument that can be 'played' with a torch or with shadows. It makes an excellent toy, and can provide lots of amusement at children's parties.

IC1 is a type 555 'timer' IC, and is connected as a free-running or astable multivibrator that produces a square-wave output signal in the speaker. The oscillation frequency is determined by C2 and by the total value of resistance appearing between pins 6 and 7 of the IC. The minimum value of this reistance is determined by R3 when the LDR is short-circuited, and the maximum value is determined by R4 when the LDR is open-circuited: the intermediate values are determined by the resistance of the LDR itself, and this is determined by the level of illumination falling on the face of the LDR.

The speaker used in the circuit can have any impedance in the range 3RO to 25R, the latter giving the highest output sound level.

Sustain Unit

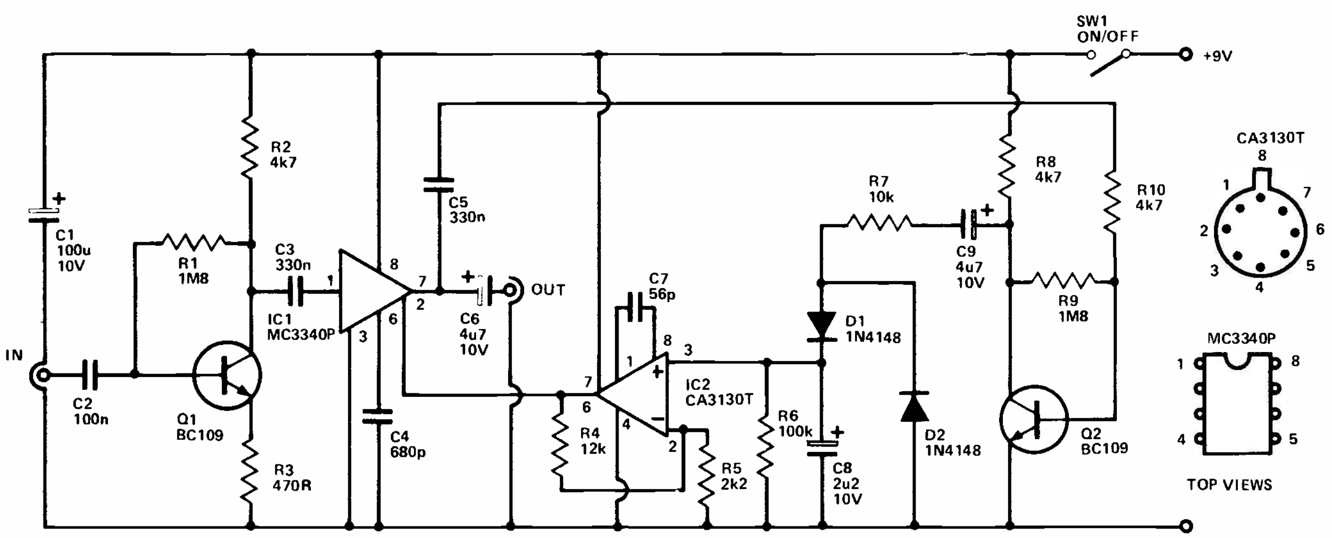
The most simple form of sustain unit is a clipping amplifier, but these inevitably introduce quite large amounts of distortion. A better method, and the one used in this unit, is to use a compression circuit having fast attack and decay times.

This type of circuit is basically a voltage controlled amplifier, the gain of the circuit being controlled by an output level sensing circuit which varies the gain to produce a fairly consistent output level. Little distortion is produced using this method.

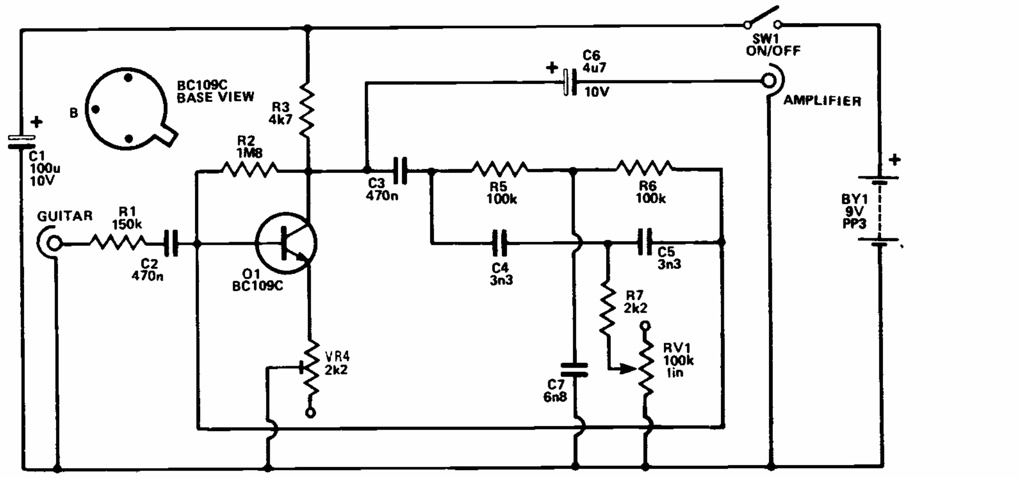
Q1 is used as a low noise preamplifier having a voltage gain of about 20dB. Its output is fed by C3 to the input of IC1, the voltage controlled amplifier device. It has a quiescent voltage gain of about 13dB, but this can be reduced to an attenuation of over 70 dB by taking pin 2 of the device several volts positive. C6 couples the output from IC1 to the output socket, and C5 couples the output to a common emitter amplifier based on Q2. The amplified signal at Q2 collector couples via C9 and R7 to a conventional smoothing and rectifier network. The positive bias produced by this network is fed to the control input of IC1 via a low gain amplifier and buffer stage based on IC2.

With low input levels (below about 1mV) the control signal is too small to affect the gain of IC1. Higher level signals produce a proportionately larger control voltage and lower gain through IC1, preventing the output level from rising much above about 30mV rms giving the required virtually constant output level. The attack and decay times of the circuit are both quite short so that the unit responds suitably rapidly to changes in input level, but neither of these time constants are so short as to cause serious distoriton.

The unit will be most effective with the volume control on the guitar set at maximum, unless the output should then be so high as to overload the unit and cause distortion.



Waa-Waa



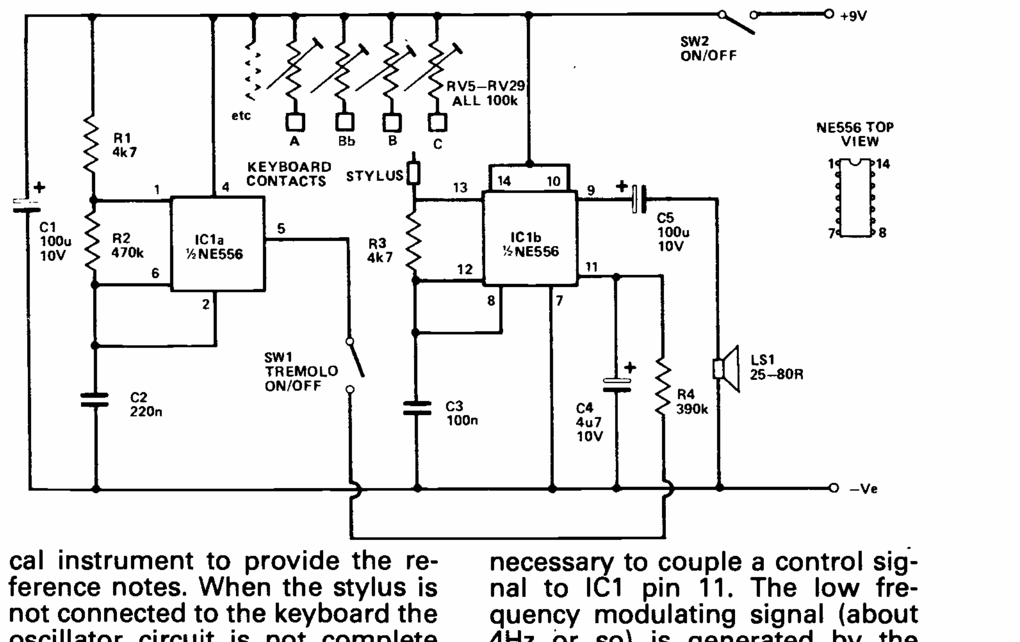
The popular Waa-Waa effect is produced using a filter which boosts a narrow band of frequencies. The position of the boosted band within the audio frequency spectrum is varied up and down, sometimes automatically but usually manually, as in the case of the design featured here. Normally a manual Waa-Waa is controlled using a pedal, but the mechanics of the unit are left to the ingenuity of individual constructors.

The circuit is based on a twin T filter; one of the T shaped networks being comprised of R5, R6, C7, and the other one consisting of C4, C5, RV1. This type of filter gives a narrow rejection notch of extremely high attenuation. This is, of course, the exact opposite of what is required! The twin T filter is made to give the required effect by connecting it between the input and output of a conventional common emitter stage based on Q1. Negative feedback is applied through the twin T filter, thus reducing the gain of the amplifier to a low level; however, there is little or no feedback at or close to the operating frequency of the network, and at these frequencies the amplifier exhibits its full gain, or very nearly so. Thus the required boost to a narrow range of frequencies is obtained.

By varying the value of RV1 it is possible to sweep the boosted band of frequencies backwards and forwards over a large part of the audio spectrum. Ideally all three resistances in the twin T network should be varied together, but this is not really practical. The circuit seems to work quite well in this simplified form.

RV4 controls the gain of the amplifier, and if this is set too low the circuit will break into oscillation at some settings of RV1. It should be set just high enough to prevent this from happening. If this gives an effect which is too sharp, R4 can be backed off a little further.

Single IC Organ



This single monophonic organ (ie it can produce only one note at a time) is intended to be used with a stylus operated keyboard. This can be made using printed circuit techniques, or it can be made from stripboard etc. if preferred. In fact anything that will provide the necessary 25 connection points will do. The organ covers two octaves including semitones; one either side of Middle C. It utilises a 556 dual timer IC which contains two 555 type timer circuits, both of which are used in the astable (oscillator) mode in this circuit.

The right hand section of the circuit is the actual tone generator, and its output feeds a miniature speaker via DC blocking capacitor C5. The frequency of operation is determined by C3, R3 and whichever of the 25 presets (RV5 to RV29) is selected using the stylus and keyboard.

Each preset is tuned to a different note ranging from C below Middle C (130.81Hz) at RV5 to the C above Middle C (523.25Hz) at RV29. The presets are adjusted to the correct pitch by aural means, using pitch pipes or a tuned musical instrument to provide the reference notes. When the stylus is not connected to the keyboard the oscillator circuit is not complete and no output is produced.

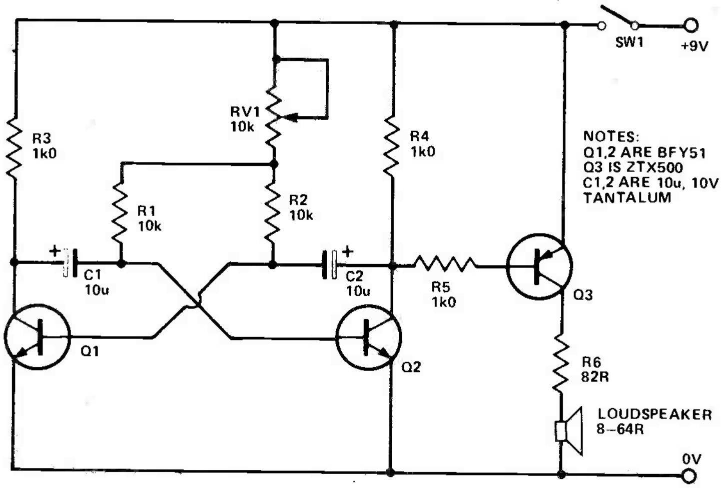
The tone generator produces a straighforward rectangular waveform which is not particularly musical. Results can be considerably enhanced by frequency modulating the tone generator to produce a tremolo effect and a richer sounding output. It is an easy matter to do this and it is merely necessary to couple a control signal to IC1 pin 11. The low frequency modulating signal (about 4Hz or so) is generated by the other section of the 556. This has its output coupled to the tone generator via R4 and tremolo on/off switch SW1. The squarewave output of IC1a would give a rather abrupt tremelo effect and so C4 is used to filter out the high frequency harmonics so as to give a much smoother and pleasant tremolo effect.

Metronome (2)

This simple circuit produces a tick-tock sound, through the speaker, the speed of which can be varied by adjustment of RV1.

Transistors Q1 and Q2 are connected in a standard astable multivibrator circuit. Potentiometer RV1 controls the charge rates of capacitors C1 and C2 and thus the operating frequency. Transistor Q3 amplifies the pulse produced by the astable and drives the loudspeaker.

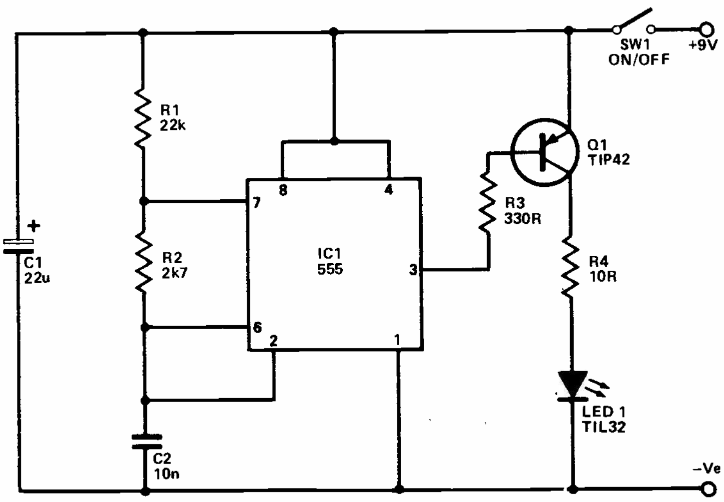
Most speaker impedances will suit the circuit although higher impedance types may not give much volume. In this case reduce the value of resistor R6 to about 27 R.



Infra-Red Intruder Alarm

This alarm sounds when an invisible beam of infra-red light is broken by the intruder. The circuit is intended to feed an existing burglar alarm system. The system is very simple, and requires no special lenses or filters for operation over distances of about 2 metres or less. A short range is usually sufficient to cover entrances to rooms, span corridors, etc.

The unit consists of two separate circuits, one to produce an infra-red beam, and the other to detect it and sound the alarm if there is a break in the signal. In common with most systems of this type, a pulsed infra-red beam is used. A modulated beam can easily be differentiated from the ambient infra-red radiation, enabling a low power beam to be employed.

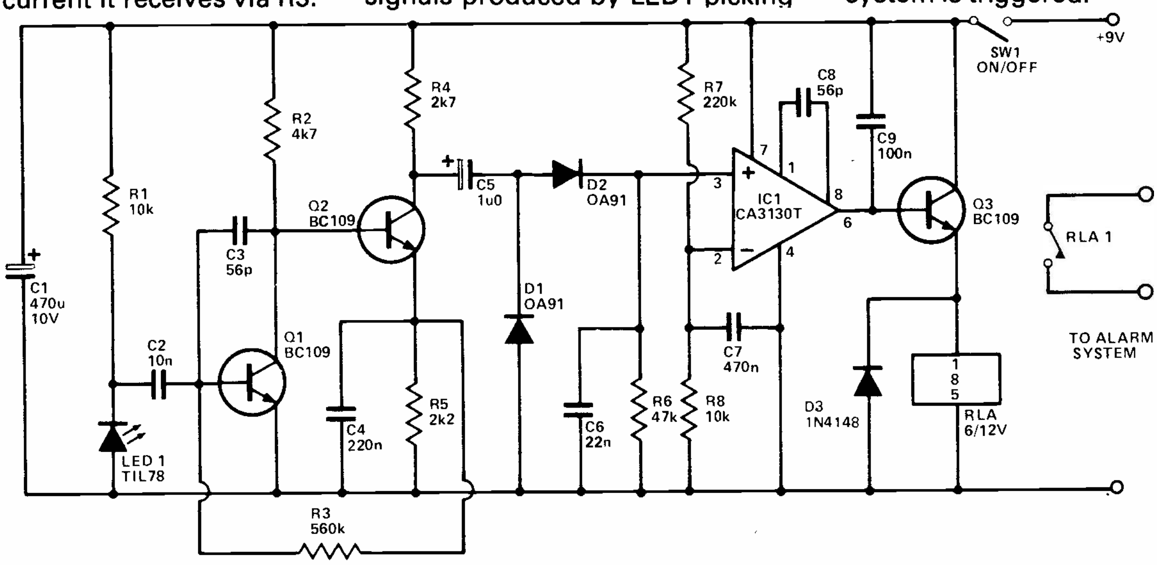


The transmitter is based on the popular 555 timer IC and it is used in the standard astable mode. R1, R2, and C2 are the timing components and these give an operating frequency of roughly 5k25Hz. A standard 555 oscillator does not produce a true squarewave output, as the output is in the high state for a longer period than it is in the low state due to the fact that the output goes high during the time that C2 charges up through the relatively high resistance of R1 plus R2 and low while C2 discharges through the low resistance of R2 and an internal transistor of IC1. The values used in this case give an output which is in the low state for only about 10% of the time.

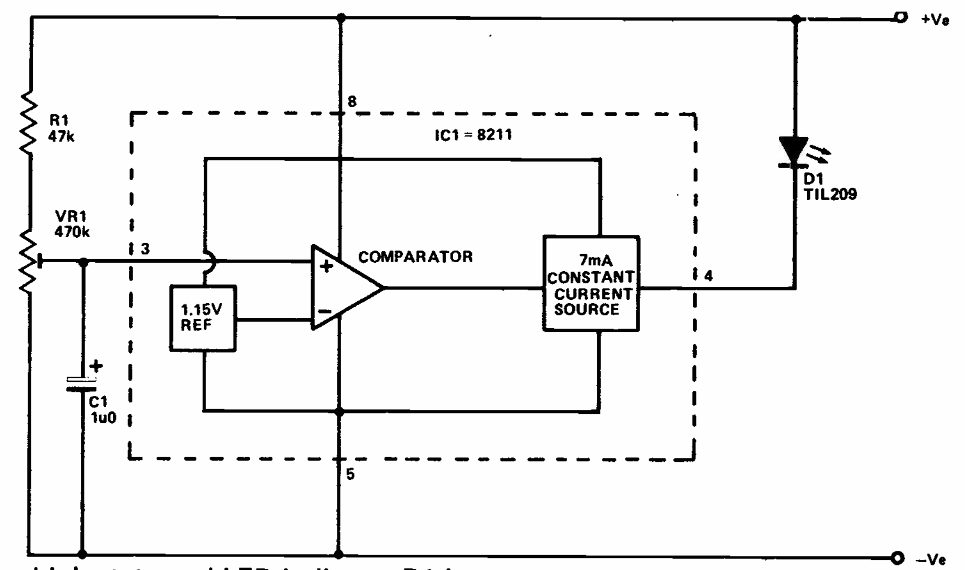
During these brief negative output excursions Q1 is pulsed on by the base current it receives via R3. It then passes a current of about 500mA, to infra-red LED1 by way of current limiting resistor R4. Of course, the average current through LED1 is only about 50 mA and this system therefore gives fairly strong pulses of infra-red together with a reasonably low average current drain. There is no output from LED1 in the visible light spectrum.

At the receiver the pulses of infra-red are detected by photodiode LED1. This is fed from the supply rails via load resistor R1. The pulses of infra-red cause the leakage current through LED1 to increase momentarily, giving a series of small voltage pulses at the junction of R1 and LED1. These are fed by C2 to the input of a straight forward high gain amplifier which uses Q1 and Q2 in a conventional two stage direct coupled arrangement. C2 and C4 are purposely given low values so that the circuit has a poor response at low frequencies. This gives good rejection of 50 Hertz signals produced by LED1 picking up the infra red radiation from mains powered lighting. The circuit has high gain at the much higher operating frequency of the transmitter circuit, though.

The output from the amplifier is coupled by C5 to a rectifier and smoothing circuit which utilises D1, D2, C6, and R6. The positive bias produced by this circuit is fed to one input of operational amplifier IC1. The other input is fed with a bias voltage by R7 and R8. Normally the fixed bias to the inverting input will be higher than the bias fed to the non-inverting input. IC1 is used as a comparator, and under these conditions its output goes high, energising the relay coil with discrete buffer stage Q3 being used to provide the fairly high drive current required. If an intruder breaks the beam momentarily, the charge on C6 rapidly decays, taking the non-inverting input of IC1 below the inverting input. IC1's output then goes low, the relay is switched off, contacts RLA1 open and the main alarm system is triggered.



Low Battery Voltage Indicator

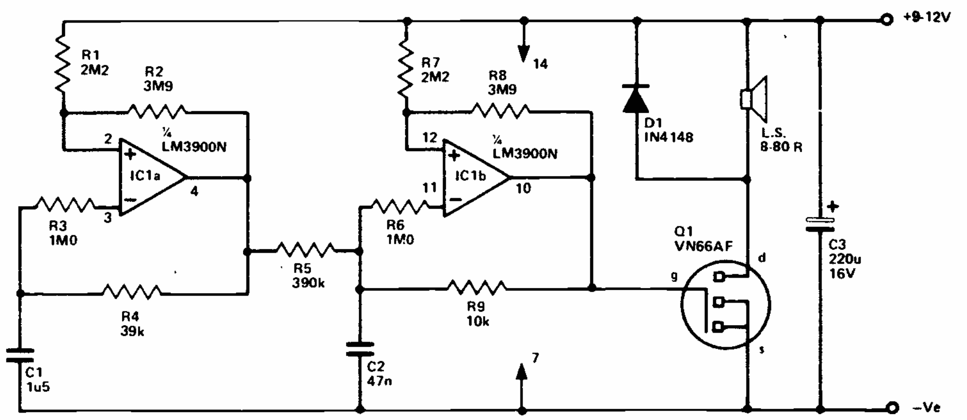


This circuit can be used to monitor a supply voltage of between about 5 and 25 volts (30V absolute maximum) and will switch on a warning light if the supply falls below some predetermined threshold level.

The circuit is actually quite sophisticated, giving good reliability and precision. This is due to the use of an 8112 voltage detector IC. A comparator forms the heart of the device, and a highly stable internally generated reference voltage is fed to the inverting input of the comparator. Its non-inverting input is available at pin 3, and in this circuit it is fed from the supply lines via the potential divider circuit which consists of R1 and VR1. The output of the comparator is available at pin 4, by way of a constant current generator which limits the output current to a nominal figure of 7 mA.

If the voltage at the non-inverting input exceeds the reference voltage, the output assumes the high state and LED indicator D1 is not switched on. If the non-inverting input voltage falls below the reference level, the output then goes low and power is applied to D1. The constant current source limits the LED current to a suitable level. In practice VR1 is adjusted so that with the supply voltage at its minimum acceptable level the non-inverting input is at a potential just marginally higher than the reference voltage. A fall in supply voltage below the threshold level then takes the non-inverting input below the reference voltage and switches on the warning light. C1 decouples any stray pick-up wich could otherwise cause spurious triggering of the circuit. The quiescent current consumption is typically about 50 uA.

Two Tone Alarm



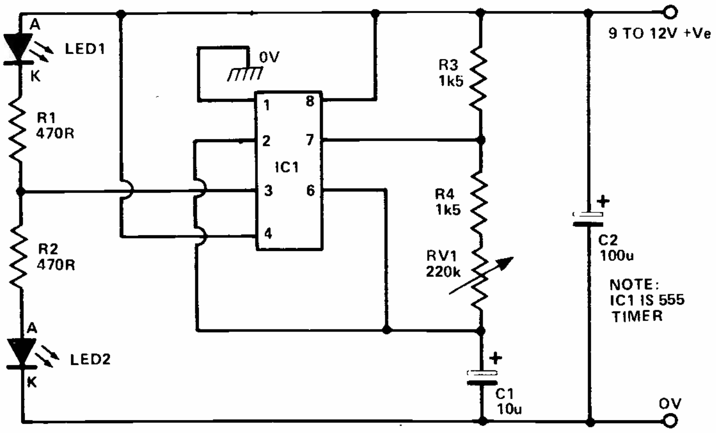
This circuit generates a penetrating two tone alarm signal having an output power of between about 250mW and 4W rms depending on the speaker impedance and supply voltage used.

The circuit is based on two of the four Norton amplifiers contained in the LM3900N IC. A Norton amplifier is in many ways similar to an ordinary operational amplifier, but it is the comparative input currents rather than the input voltages that determine the output voltage. IC1b is used in a type of relaxation oscillator which generates the audio tone. Initially the bias current flowing into the non-inverting input takes the output high, and C2 begins to charge via R9. This causes the current flowing into the inverting input through R6 to gradually increase as the voltage on C2 builds up, until it exceeds the non-inverting input bias current. The output of IC1b then goes low, causing C2 to discharge through R9 until the inverting input current becomes less than that flowing into the non-inverting input. IC1b output then goes high, and the procedure starts again from the beginning. Note that when IC1b output went from the high to the low state, this resulted in R8 draining off some of the non-inverting input bias current where it had previously added to it. This makes it necessary for C2 to discharge considerably before the current into the inverting input drops below that applied to the non-inverting input. This effect is a form of "hysteresis", and is essential to the operation of the circuit. R8 also provides positive feedback which ensures that once IC1b output starts to change polarity, it rapidly and reliably switches from one state to the other.

The squarewave output of IC1b is at quite a high impedance, and so the loudspeaker is driven by way of a common source amplifier using VMOS transistor Q1. IC1a is used in a second oscillator circuit, but this has component values which give oscillation at a frequency of only a few Hertz. Its output is loosely coupled to C2 by R5, and it frequency modulates the tone generator to produce a sort of warbling effect. This gives a much more noticeable and less easily masked signal than a straight forward audio tone.

Indicators and Alarms

LED Flasher



This circuit can be used to either pulse a single LED on and off repetitively or to similarly drive a pair of LEDs in anti-phase, so that one LED turns off when the other turns on, and vice versa. In either case, the flashing rate of each LED is variable from about 15 flashes per minute to 2,000 flashes per minute via a small pre-set pot.

The circuit can be used to add visual interest to a variety of toys, gadgets, and istruments. Model railway enthusiasts can use the unit to simulate flashing lights on miniature police cars and ambulances, etc, or to simulate warning beacons on Zebra crossings.

IC1 is a type 555 'timer' IC, and is connected as a free-running or astable multivibrator that produces a square-wave output signal at pin 3. When this output signal is high it cuts LED 1 off and drives LED 2 on, and when it is low it pulls LED 1 on and cuts LED 2 off; the two LEDs thus turn on and off in anti-phase.

The operating frequency of IC1, and thus the flashing rate of the LEDs, is determined by the values of C1 and R4-RV1: the flashing rate is variable between roughly 15 and 2,000 flashes or cycles per minute via RV1. The ON currents of the LEDs are limited to safe values by the R1 and R2 470R limiting resistors.

The flashing rate can be carried via RV1. If you want the circuit to operate with only one flashing LED, you can either short out the unwanted LED or remove it and it's associated 470R resistor from the circuit.

Light Change Detector

This simple light activated switch can be employed in intruder alarm systems or in certain other applications where a proximity detector is required. It is not a light operated switch of the type which responds to some particular ambient light level but instead it responds to rapid changes in light level. It will for example, trigger if a torch is shone on or near the photocell or if someone passing in front of the unit casts a shadow onto the photocell. The unit is unaffected by natural, slow changes in light level.

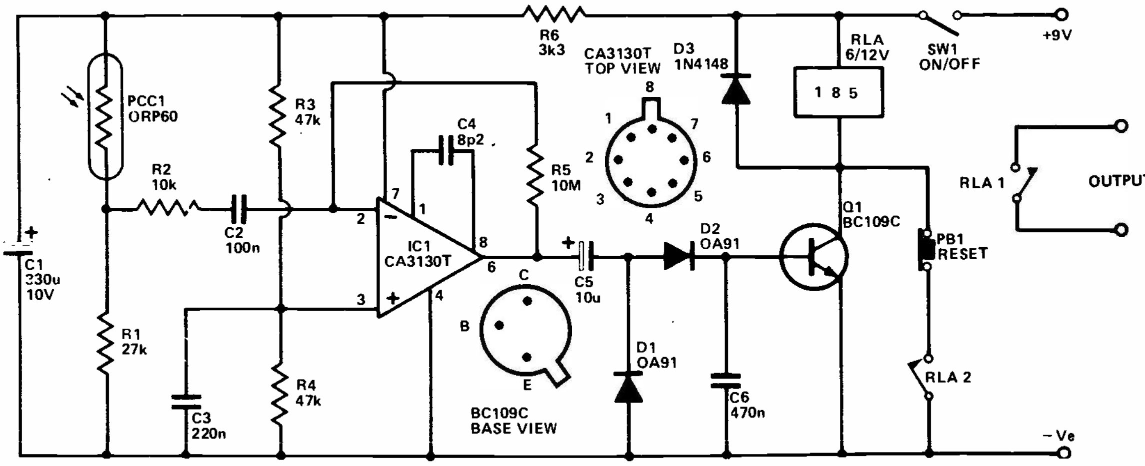
The photocell used in the unit is cadmium sulphide photoresistor, PCC1, and together with R1 this forms a potential divider connected across the supply lines. The voltage at the junction of R1 and PCC1 depends upon the resistance of PCC1, which in turn depends upon the light level.

IC1 is an operational amplifier used in the inverting mode. R2 and R5 form a negative feedback network which set the voltage gain of the circuit; high gain is required in order to give the unit good sensitivity.

The output from PCC1 and R1 is coupled to the input of the amplifier, but C2 provides DC blocking so that the DC output from these components is of no consequence. However, rapid changes in the output voltage from the photocell circuit will be passed to the input of the amplifier and will appear at the output.

The output from IC1 is coupled by C5 to a smoothing and rectifier circuit consisting of D1, D2, and C6. When the unit is activated, the positive bias produced across C6 is sufficient to bias Q1 into conduction so that it energises the relay coil which forms its collector load. RLA2 then closes and maintains the supply to the relay coil, so that the circuit latches on. The alarm or other controlled equipment is operated by RLA1.

SW2 can be used to break the supply to the relay and thus reset the circuit. SW1 is the on/off switch; C4 is the compensation capacitor for IC1 and D3 is the normal protective diode.



Opto-Thermo Alarm

This useful project can be powered from a 9V to 12V supply, and produces a pulsed-tone alarm signal in a small speaker when light or temperature levels go beyond pre-set limits. The unit can be made to activate either when these levels go above or fall below pre-set values, depending on the manner in which the input sensors (an LDR or light-dependent resistor for 'light' operation, or a thermistor for 'temperature' operation) are connected to the unit.

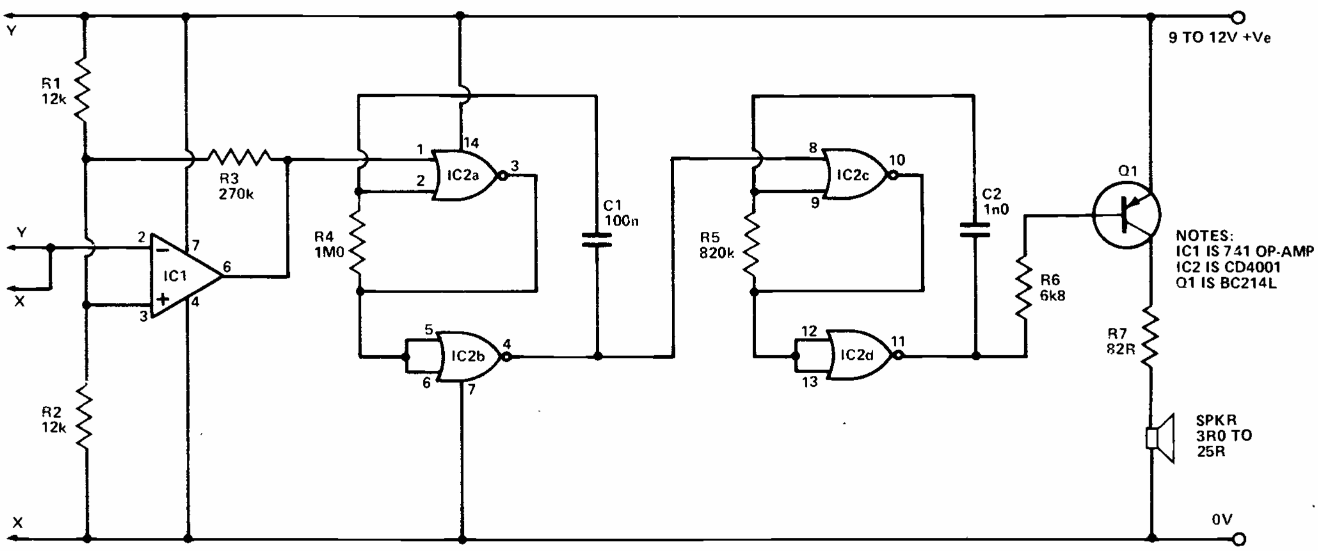
The unit has a variety of uses in the house and in the car. In the car, it can be used to give a warning of road ice or of engine or gearbox overheating. In the home, it can be used to give a warning of a burnt-out night-light or a failed heating system in a child's room, or it can be used as a 'dawn' alarm.

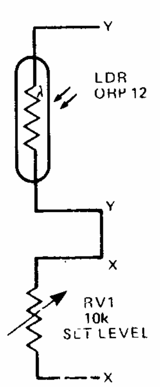
IC1 is a type 741 operational amplifier wired as a volatage comparator with a small amount of regenerative feedback. A fixed 'half supply' reference voltage is fed to input pin 3 of the op-amp via R1 and R2, and a variable voltage is fed to pin 2 via RV1 and the LDR or thermistor. The output of IC1 is used to activate (turn on or off) a 'slow' gated astable multivibrator formed by IC2a and IC2b, and the output of IC2b is used to activate a 'fast' gated astable formed by IC2c and IC2d which has its output fed to the speaker via Q1.

When the pin 2 voltage of IC1 is below that of pin 3, the output of IC1 is high, and the two astables are gated off and no output is produced from the unit. When the pin 2 voltage of IC1 is above that of pin 3, the output of IC1 is low, so slow astable IC2a-IC2b is gated on and its output alternately switches the fast IC2c-IC2d astable on and off to produce a pulsed-tone in the speaker.

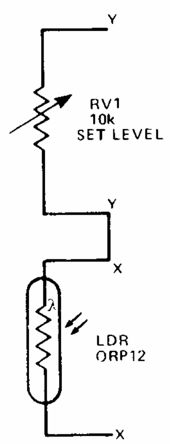
Connect the input sensor (either an LDR or a thermistor) and RV1 to the inputs of the unit to obtain the desired types of operation. For light operation, the LDR must present a resistance in the range 900R to 9kR at the desired trigger level: an ORP12 is suitable for use in most cases. For temperature operation, the thermistor must be a negative-temperature-coefficient (NTC) type that presents a resistance in the range 900R to 9kR at the desired trigger level: a VA1066S is suitable for use in most cases. The speaker can have any impedance in the range 3R to 25R, the latter value being preferred.

The unit can be powered from any DC supply in the 9V to 12V range. In use RV1 is simply adjusted so that the alarm just activates at the desired light or temperature level. The unit produces an attractive pulsed-tone signal when it is activated.

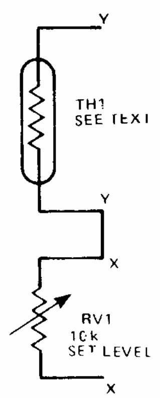




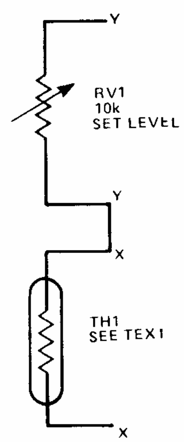
Dark-operated



Light-operated



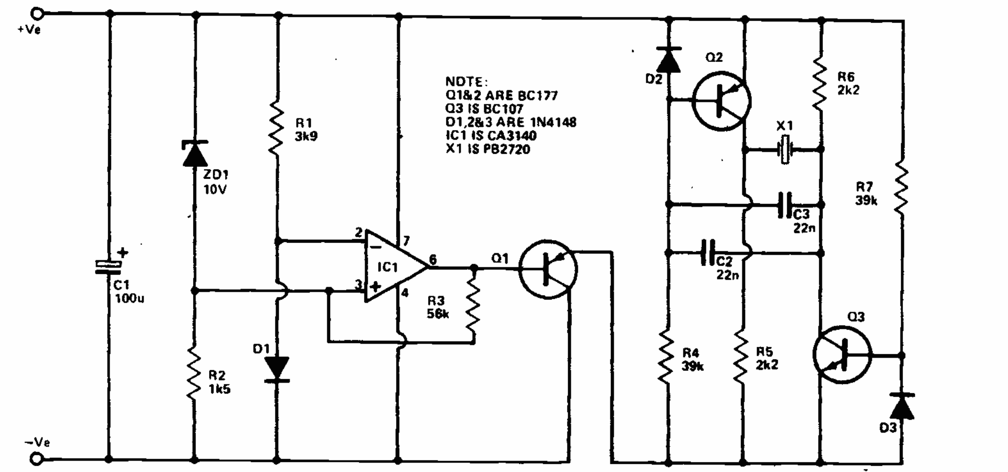
Under-temperature



Over-temperature

Indicators and Alarms

Car Battery Alarm



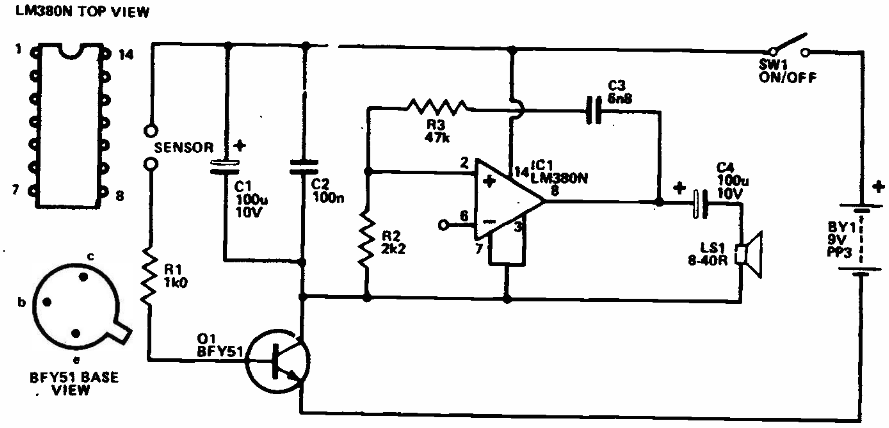
This circuit is intended for use with a 12V car or boat battery and sounds an alarm if the battery voltage drops below about 10V. An audible alarm is used in preference to the more usual indicator light because an audible signal is far less likely to be missed.

IC1 and its associated circuitry form a voltage detector circuit; IC1 is used here as a voltage comparator rather than an operational amplifier. R1 and D1 provide a stabilised potential of about 0V7 to the inverting input of IC1. The voltage supplied to the non-inverting input is equal to the supply voltage minus the nominal 10V dropped through zener diode ZD1. Thus the potential at the non-inverting input will normally be about 2V or more and since this is higher than the voltage fed to the other input the output assumes the high state. Therefore Q1, an emitter follower buffer stage, becomes cut off and no power is fed to the audio alarm circuit at its output.

If the supply voltage falls below about 10V7, then the voltage fed to the non-inverting input falls below the reference level at the other input and IC1's output goes low. Q1 is then switched on and power is connected to the alarm circuit fed from its output. Due to imperfections in the performance of practical zener diodes, the actual voltage at which the alarm is triggered is a little lower than the theoretical one and is typically a fraction over 10V (or 11V if preferred, by using an 11V component in the ZD1 position). R3 is used to introduce a small amount of positive feedback which ensures that the unit switches cleanly from one state to the other and prevents erratic operation.

The alarm generator circuit uses Q2 and Q3 as an astable multivibrator and these drive ceramic resonator X1 with the anti-phase signals at their collectors. This gives the necessary large voltage swing to X1 and produces a reasonably loud output at about 2kHz or so.

Water Alarm



This simple water detector circuit can be used in such applications as a rain alarm, cistern overflow alarm, or just to indicate when the water in a bath has reached the required depth.

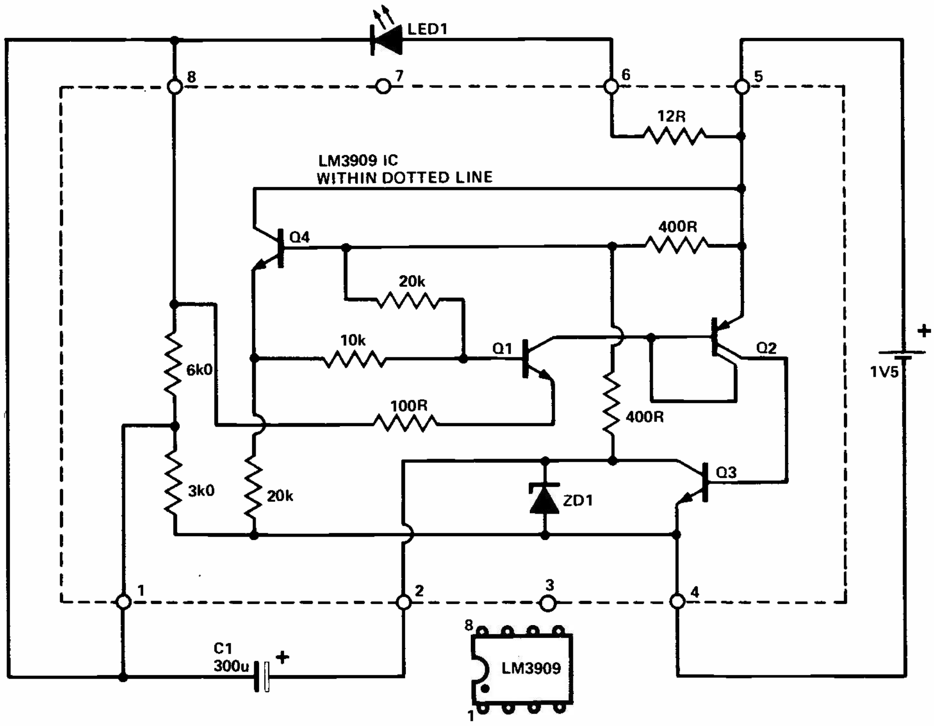
The circuit consists of two sections; an electronic switch and an audio alarm generator: Q1 is used as the switch and under normal conditions it is cut off, supplying no significant current to the alarm generator circuit which forms its collector load. The sensor in Q1's base circuit merely consists of two pieces of metal insulated from one another, arranged so that when the rain, bath water, or whatever touches the sensor, it bridges the two pieces of metal. Although pure water is a very poor conductor of electricity, tap water is likely to contain amounts of impurities which will be sufficient to make the water conduct reasonably well. Thus when water is detected by the sensor, its resistance falls to a relatively low level (typically a few kilohms), biasing Q1 hard into conduction. Virtually the full supply voltage is then supplied to the alarm circuit, and the alarm sounds. R1 is a current limiting resistor which prevents Q1 from passing an excessive base current if a short circuit or very low impedance should appear across the sensor terminals.

An LM380N audio power amplifier device (IC1) is used to generate the audio alarm signal. It is made to oscillate by using C3, R3, and R2 to give frequency selective positive feedback between the output and non-inverting input. The circuit oscillates at approximately 600 Hz, and provides an output of a few hundred milliwatts to a miniature loudspeaker. This gives a reasonably loud and penetrating sound.

Under quiescent conditions the circuit consumes less than one microamp, and so even with continuous use the battery will have virtually its shelf life. With the alarm sounding, the current consumption increases to about 30 to 100mA, (depending on the speaker impedance). Note that C2 should be mounted physically close to IC1.

A little ingenuity must be used when making the sensor. In most applications a small piece of stripboard will be all that is needed.

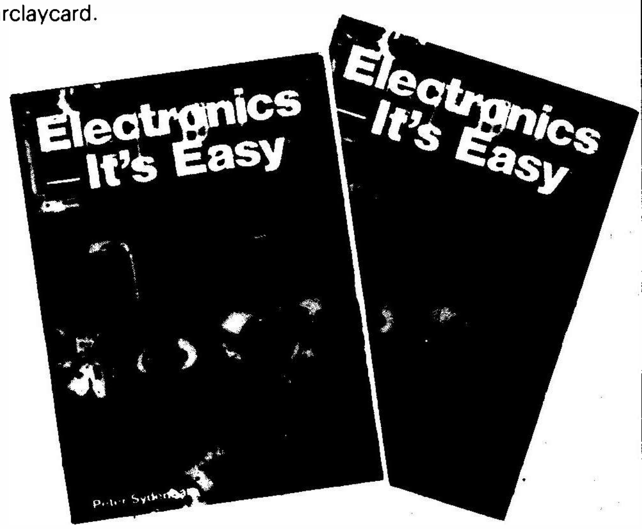
Low-Power LED Flasher



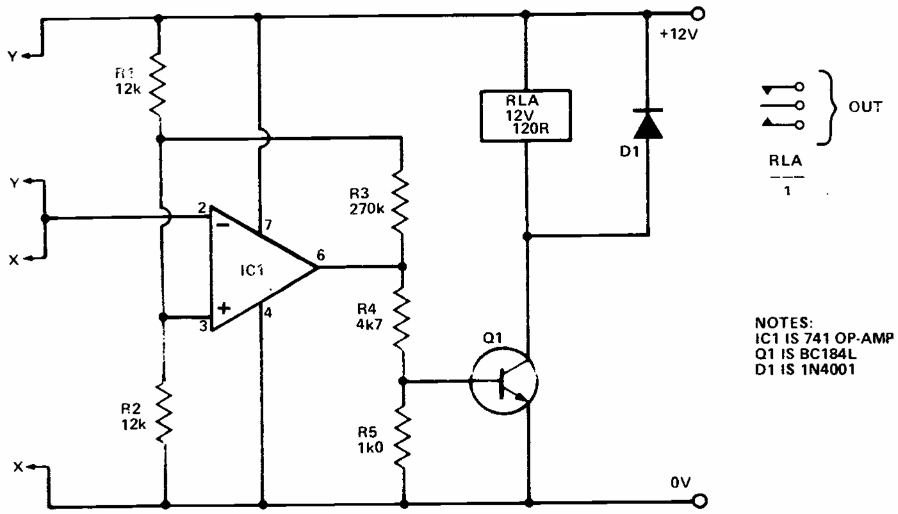
National Semiconductor have produced an integrated circuit to be used specifically for flashing an LED, even at voltages as low as 1V1, with an average current consumption orders of magnitude below that of an LED on its own.

The circuit achieves it's very low current consumption because the LED is only illuminated 1% of the time, and only Q4 is turned on for the rest of the time – drawing a current of only 50uA while on. The 300u capacitor determines the flash rate by charging up via the two internal 400R resistors and the 3k resistor. Q1 and Q2 are turned off until the voltage at the positive end of C1 reaches about 1V; the exact voltage is determined by the junction voltage drop of Q1 and Q4 plus the voltage divider across Q4's base and emitter.

When the voltage at pin 1 is 1V more negative than the positive supply (pin 5), Q1 starts to turn on. This in turn switches on Q2 and Q3. Q3 is a medium-power transistor that can hadle 100mA and rapidly brings pin 2 close to zero volts. As the capacitor has a charge it makes terminal 1 (the negative end of the capacitor) go below supply zero. C1 then supplies a high current pulse to the LED, limited by the 12R resistor between pins 5 and 6. The cycle then repeats itself.



Opto-Thermo Switch



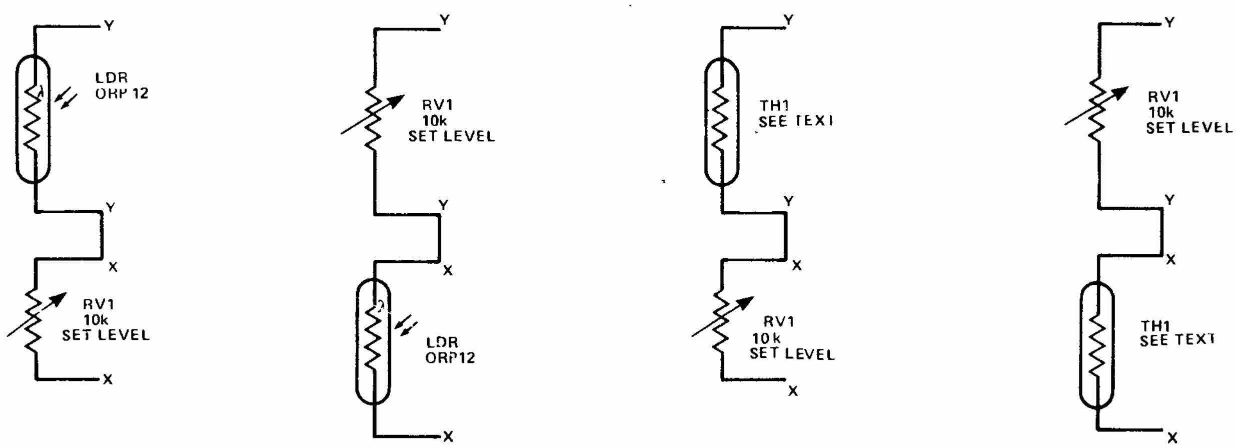
This simple but highly versatile unit has a relay output that can be activated by either optical (light) or thermal (temperature) levels. The unit can be made to activate either when these levels go above or fall below pre-set values, depending on the manner in which the input sensors (an LDR or light-dependent resistor for photo operation, or a thermistor for thermal operation) are connected to the unit. The unit can thus function as either a 'brightness' switch, a 'darkness' switch, an over-temperature switch, or an under-temperature switch.

The unit has a multitude or practical uses. In the opto mode it can be used to automatically turn on lights when darkness falls, or to activate an alarm if a light is shone into a normally-dark area such as a cupboard or safe. In the thermo mode it can be used to turn on heating when the temperature falls below a pre-set value, or to activate a cooling system or sound an alarm when the temperature rises above a pre-set value.

IC1 is a type 741 operational amplifier that is wired as a voltage comparator. A fixed 'half supply' reference voltage is fed to input pin 3 of the op-amp via R1 and R2, and a variable voltage is fed to input pin 2 via RV1 and the LDR or thermistor. The circuit action is such that the op-amp output is normally low and Q1 and the relay are off, but the output abruptly switches high and drives Q1 and the relay on when the pin 2 voltage falls below the pin 3 voltage. R3 introduces a small amount of hysteresis so that the circuit switches sharply and switches at slightly different ON and OFF levels, thus eliminating relay 'chatter' problems. D1 suppresses back-emfs from the relay coil, and this protects Q1 against damage from this source.

Connect the input sensor, either an LDR or a thermistor (and RV1) to the inputs of the unit to obtain the desired type of operation. For opto operation, the LDR must present a resistance in the range 900R to 9k at the desired trigger level an ORP12 is suitable for use in most cases. For thermo operation, the thermistor must be a negative-temperature coefficient (NTC) type that presents a resistance on the range 900R to 9k at the desired trigger level: a VA1066S is suitable for use in most cases. The relay can be any 12 volt type with a coil resistance greater than 120R.

The unit must be powered from a 12 volt supply. In use RV1 is simply adjusted so that the relay just activates at the desired light or temperature level. External circuits can be controlled via the relay contacts. Take care to connect D1, Q1, and IC1 in the polarity shown.



Dark-operated Light-operated Under-temperature Over-temperature

Differential Temperature Switch

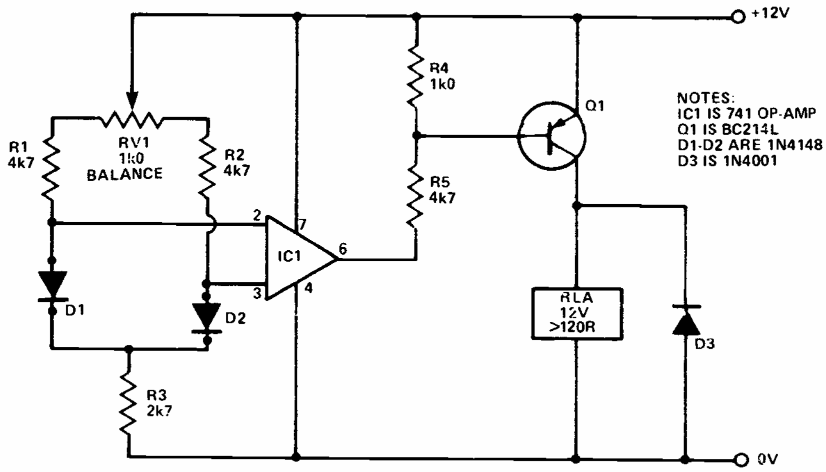
This inexpensive unit can form the basis of a number of sophisticated household-control systems. The circuit uses a couple of ordinary silicon diodes as temperature-sensing elements, and uses a relay as an output 'switch.' The circuit action is such that the relay turns on only when temperature 'A' (sensed by D1) is higher than temperature 'B' (sensed by D2), and this action occurs irrespective of the absolute value of either temperature. The circuit action can be effectively reversed, so that the relay turns on only when temperature 'A' is below that of temperature 'B', by simply transposting the measurement designations of D1 and D2.

Ordinary silicon diodes can develop forward voltages of several hundred millivolts at current levels of the order of 1 mA, the precise voltage value depending on the value of current and the characteristics of the individual diode that is used. All silicon diodes, however, have a virtually identical temperature coefficient of about −2mV/°C, and can thus be used as accurate temperature-indicating devices.

In the circuit the two temperature-sensing diodes (D1 and D2) have currents passed through them via the RV1-R1-R2-R3 network; RV1 allows the relative values of the two currents to be adjusted over a limited range so that the diodes produce almost identical forward voltages when they are both at the same temperature. Consequently, the differential or 'difference' voltage between the two diodes is directly proportional to the difference in their temperatures.

This difference voltage is fed to the input terminals of the IC1 operational amplifier, which is connected as a voltage comparator or differential voltage switch, and the output of the op-amp is fed to the relay via Q1. The action is such that the relay turns on when the temperature of D1 rises above that of D2.

The unit must be powered from



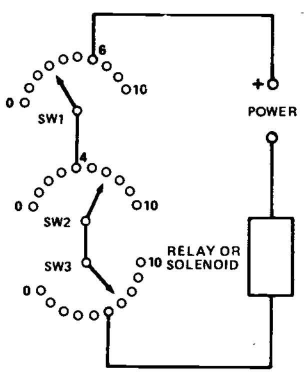
a 12 volt supply. In use, RV1 is simply adjusted so that the relay is just off when both sensing diodes (D1 and D2) are at the same temperature. The relay should then turn on if the temperature of D1 is raised a small amount above that of D2: note that at normal room temperature this action can be checked by simply touching D1, so that body heat produces the required differential. External circuits can be controlled via the relay contacts. The relay can be any 12 volt type with a coil resistance greater than 120R.

This switch can, for example, be used to activate a blower motor to ensure that a cellar or basement is automatically warmed by the outside air if the external air temperature is above that of the cellar or basement. Alternatively, it can be used to activate a solenoid valve to ensure that a storage tank is automatically filled only from the hotter of two alternative water sources, etc.

Combination Lock

Most combination locks are based on the simple arrangement shown in (a). This merely consists of three ten way rotary switches wired in such a way that they will connect power to the relay or solenoid if the correct combination is set (6-4-5 in this case). This basic circuit is not often used in practice since it does not take very long to quickly adjust the switches through all the 1,000 possible combinations (0-0-0 to 9-9-9 inclusive) if it is done in a logical manner.

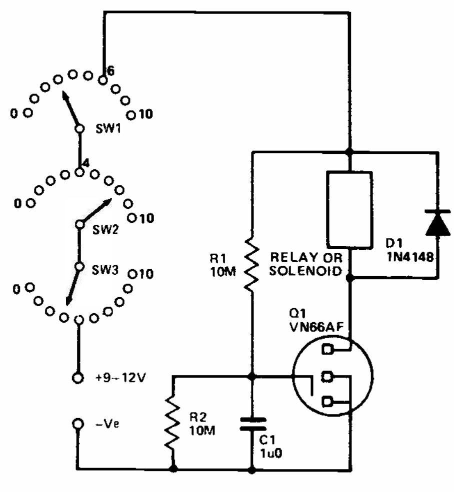
One of the simplest and best methods of overcoming this problem is simply to build a delay circuit into the unit so that power is not supplied to the solenoid or relay until the correct combination has been present for a few seconds. Quickly running through all the possible combinations is then ineffeceive at "cracking" the unit, as the delay circuit will prevent the unit from responding when the correct combination is briefly present. Anyone trying to "crack" the unit is very unlikely to succeed unless they know of the delay circuit and are prepared to devote a good deal of time to finding the correct combination.



(a)

Circuit (b) has an additional time delay circuit which is based on VMOS device Q1. When the correct combination is set on SW1 to SW3 and power is supplied to the circuit, C1 will be uncharged, giving zero gate bias to Q1. Q1 is, therefore, switched off and no significant current is supplied to the relay or solenoid which forms its drain load. C1 slowly charges via R1 and after about 4 s the voltage on C1 will be large enough to bias Q1 into conduction and switch on the relay or solenoid.

R2 discharges C1 when the unit is reset, so that it is quickly ready to operate properly once again. R2 limits the maximum gate voltage of Q1 to about half the supply voltage, but this is more than adequate to bias the device hard into conduction and is of no practical consequence. D1 suppresses the high back emf produced across the relay or solenoid when it switches off and prevents possible damage to Q1.



(b)

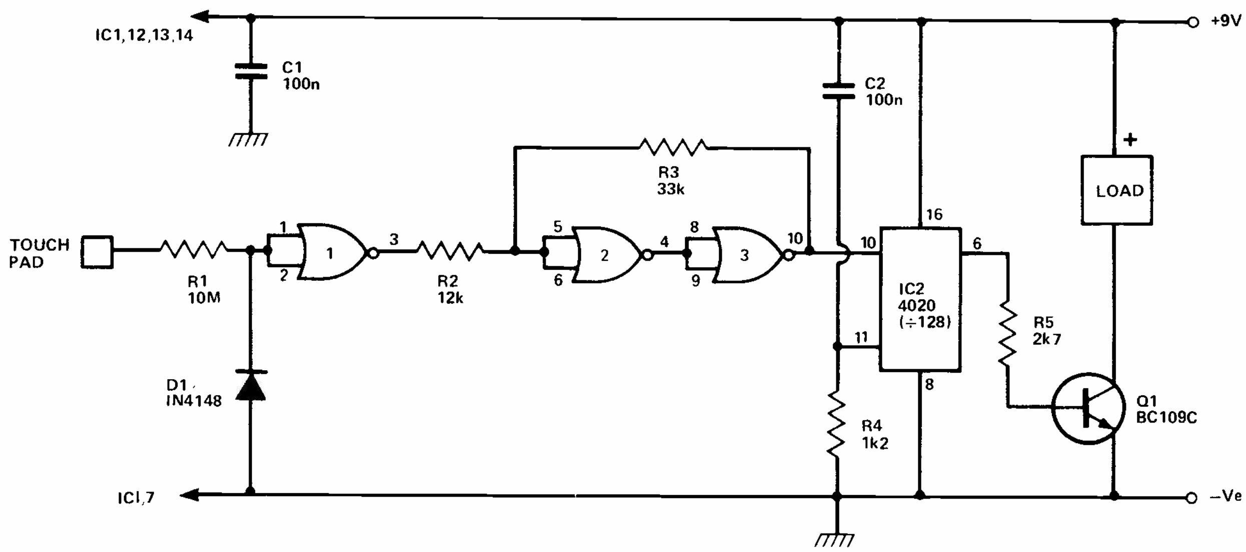
Touch Switch

This touch switch is designed to provide on/off switching for 9 volt battery operated equipment having a current consumption of up to 100mA. It has a single contact which is briefly touched in order to change from on to off or vice versa. The circuit is operated by stray pick-up of mains hum which is coupled to the input of gate 1 (which, like the other three gates employed in the unit, is connected to act as an inverter) via R1 when the input contact is touched. As IC1 is a CMOS device it has a very high input impedance, and the input signal will be capable of switching gate 1 input from one logic state to the other. The input impedance of the circuit is so high that the reverse resistance of D1 is used to tie the input to earth under quiescent conditions, so as to prevent spurious operation. R1 acts as a low pass filter in conjunction with the input capacitance of the circuit, and this attenuates high frequency noise which may be present on the 50Hz mains signal.

The output from gate 1 still contains significant noise products, and also has a rise time which is inadequate to drive the final stage of the circuit. This is overcome by using the trigger circuit based on gates 2 and 3. R3 tends to hold gate 2 input in the same state as gate 3 output, resisting any change in logic state caused by gate 1 output due to the coupling through R2. R2 has a lower value than R3, and so gate 1 can operate the trigger circuit if its output signal is of adequate amplitude. The main 50Hz signal will be strong enough, but the noise spikes will not, and are thus eliminated from the output of the trigger. Once the output of the trigger starts to change state, the coupling through R3 ensures a rapid change.

IC2 is a 14 stage binary (divide by 2) counter, and Q1 is driven from the output of the seventh stage via current limit resistor R5. C2 and R4 provide a positive reset pulse to the counter at switch on so that the outputs are low, and Q1 is switched off. The controlled equipment forms the load for Q1, and obviously receives no significant power. If the touch contact is operated, a 50Hz signal is fed to IC2 and the 7th stage output changes state every 64 pulses. As this output goes high and low the load is switched on and off. In practice the contact is touched just until the unit switches to the desired state (which one tends to do automatically).

The unit consumes about 1uA. in the "off" mode and approximately 3mA in the "on" state.



Light Sensitive Switch

Light dependent resistor LDR1, type ORP12, has the property of having a very high resistance in complete darkness (over 1MR), but in the brightest light this falls to nearly 300R (although individual devices vary greatly).

For simplicity the circuit will be described for 'dark on'.

When daylight is falling on the ORP12 the resistance is fairly low – the actual value is not important but it will be in the range of 300R to 3kR. With R1, the two components form a potential divider with the base of Q1 connected to the junction; Q1 is biased on, Q2 is held off and, following the chain of R5 and R6, it will be seen that the base of Q3 is connected to the positive supply, and is therefore non-conducting.

As the light on ORP12 decreases, the voltage at Q1 base falls; the current through Q1 fails, the voltage across it increases, and this starts to bias Q2 into conduction.

As Q1 and Q2 share the same emitter resistor there is a regenerative action. Q2, by becoming conductive, raises the voltage on the emitter side of R4 and, since the base voltage on Q1 is still at the same level, this biases Q1 even more into non-conduction.

The switching action is extremely rapid once it has started and the effect is that, although the voltage at the base of Q1 is changing gradually, Q2 is switched completely on at a certain point, and Q1 is switched off.

When Q2 is switched on, the voltage at the collector falls and Q3 is biased into full conduction through R6. With Q3 on, current is passed through the buzzer.

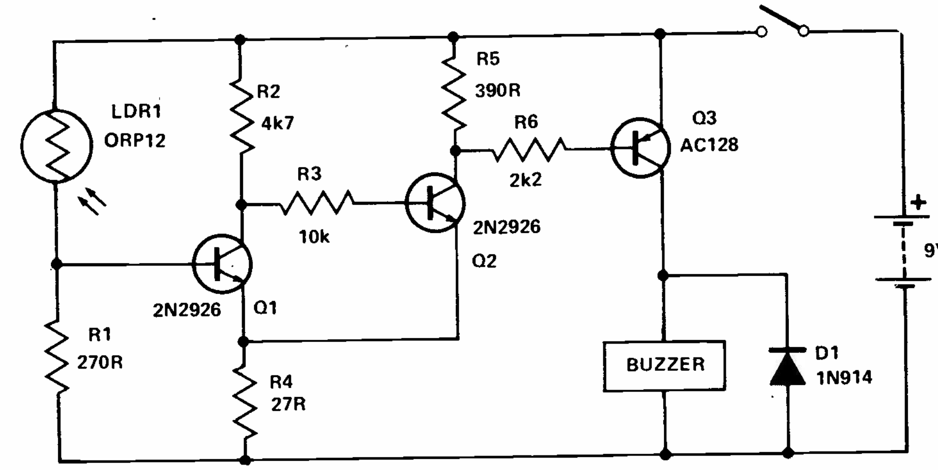
A relay can be used in place of the buzzer – its coil resistance hardly matters – however, relay coils should always be by-passed with a diode.

A 'dark off' (or 'light on') circuit can be made by reversing ORP12 and R1 – which now becomes 33k. In this condition Q1 is on but as the light level increases the voltage at the base of Q1 falls and the same action applies as before.

In both versions of the circuit R1 can be replaced by a variable resistor and this enables the light level at which the circuit triggers to be varied.

Note that Q1 and Q2 are silicon N-P-N types whereas Q3 is a germanium P-N-P type. Q3 passes quite a large amount of current and can be destroyed unless a heat sink is used – one of the fin types which clip over the body should be adequate.

When Q3 is off the current consumption is very low – about 4mA.



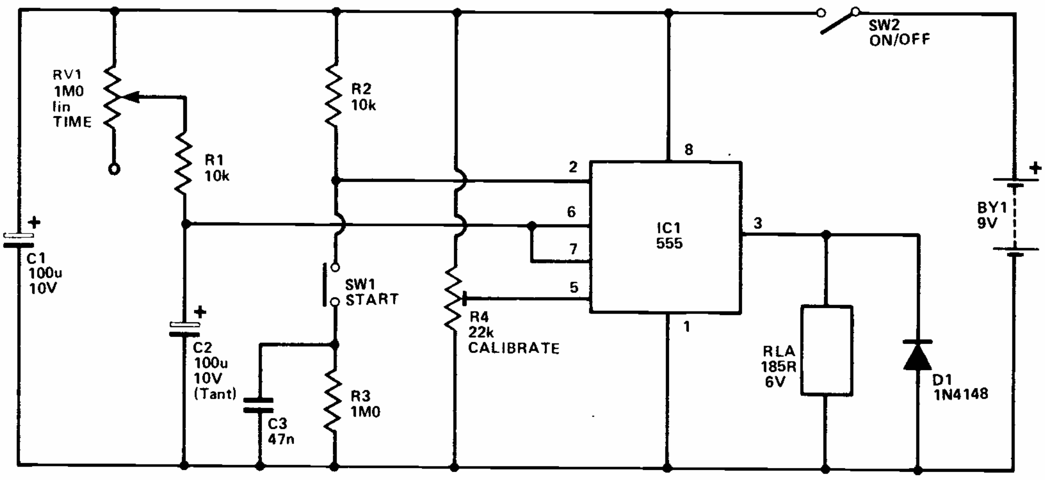
Sequence and Timing

555 IC Timer

In the monostable multivibrator mode the 555 IC requires a trigger pulse which takes the trigger input below 1/3 of the supply voltage. The normally low output then goes high for a period determined by a C-R network, the pulse length being approximately 1.1 CR seconds. However the output will stay high as long as the trigger (pin 2) is held below $1/3V^+$

This circuit is for a timer having a 1 to 100 second range, suitable for use as an enlarger timer, for example. It is triggered by depressing SW1, generating a brief negative pulse before C3 charges via R2. This makes it impossible to prolong the input pulse and produce a false output period by keeping SW1 depressed. C3 is quickly discharged by R3 when SW1 is released, rendering the trigger circuit ready for the next operation.

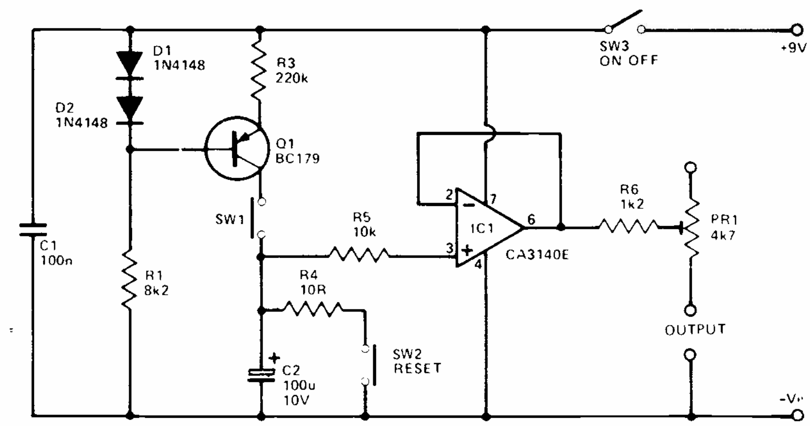
The tolerance of timing components RV1, R1, and C2 make it impossible to obtain a highly predictable time range, and this is overcome by the inclusion of R4. Normally the output pulse ends when the charge on the timing capacitor reaches 2/3 V+. R4 can be used to raise or lower the threshold voltage, to extend or shorten the output pulse, to obtain the appropriate timing range.



D.M.M. To Stopclock Converter

This simple add-on circuit can be used with a DMM switched to the 1mA range to give a stopclock having a range of 0-99s (or 0-199s for a 3½ digit instrument). It can also be used with an ordinary analogue or multimeter or panel meter, giving a range of 1-100s, but the resolution will be lower than with a digital instrument.

The unit relies on the fact that a linear rise in voltage is produced across a capacitor if it is fed with a constant charge current. The capacitor (C2) must be a high quality type. The use of a tantalum bead component is, therefore, recommended. C2 cannot simply be charged from the supply lines via a resistor, since the voltage across the resistor would drop as the voltage across C2 increases. This would give a decreasing charge current as C2 charges exponentially and the required linear voltage slope would not be produced. C2 is, therefore, charged from a conventional constant current source which is based on Q1. D1, D2 and R1 form a simple shunt regulator circuit which bias the

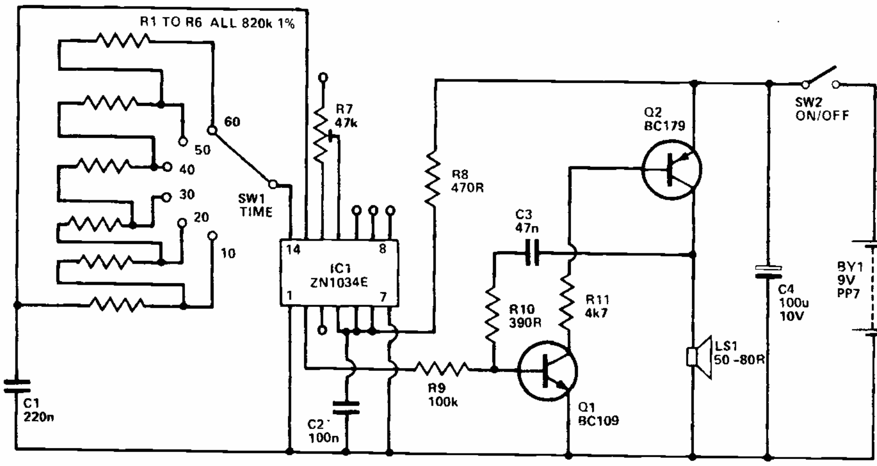


base of Q1 approximately 1V3 below the positive supply potential. There is a voltage drop of about 0V65 across the base emitter terminals of Q1, giving about 0V65 across emitter resistor R3. This gives an emitter current of roughly 3uA and, as the collector and emitter currents of a high gain device (such as the BC179 used in the Q1 position) are virtually identical, a constant charge current of about 3uA is fed to C2 when SW1 is operated. This low charge current together with the fairly high value of C2 produces a suitably long time constant.

It is essential that the voltmeter circuit takes no significant current from C2 as this would affect accuracy and would result in a decaying reading at the end of a timing run. Operational amplifier IC1 is, therefore, used as a unity gain buffer stage which gives an input impedance of about 1.5 million megohms and ensures that there is no significant loading on C2. PR1 enables the sensitivity of the voltmeter circuit to be adjusted to the correct level. In practice, SW1 is depressed for (say) 90s and then PR1 is adjusted for the appropriate reading on the DMM.

SW2 is a reset switch and this discharges C2 (via current limiting resistor R4) if it is briefly operated. SW3 is an ordinary on/off switch.

ZN1034E Timer



Long timing periods are difficult or even empossible to obtain using most simple timing circuits, such as those using the well-known 555 timer IC.

The problem is that such circuits usually provide a timing period of about RC seconds, and times of more than a few minutes require unrealistically high values for the timing resistor and capacitor.

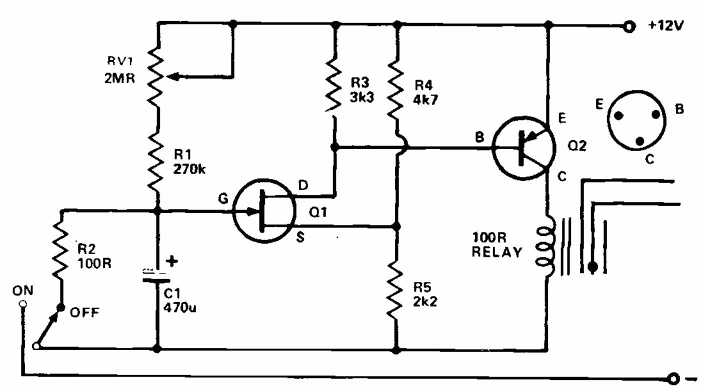
The ZN1034E device is specifically intended for applications where long times are required, and by using an oscillator, a 14 stage binary divider (divide by 4,096 circuit), and a logic control section, an output time of between about 2,700 and 7,500RC seconds can be attained! Times of a few minutes to a few hours can be obtained without having to use an electrolytic timing capacitor, giving excellent reliability and accuracy.

This simple timer circuit provides an audible alarm at the end of the timing period which can be varied from 10 to 60 minutes in 10 minute increments. C1 is the timing capacitor, and the timing resistance is formed by which ever of the series of resistors (R1 to R6) is switched into circuit using SW1. The timing resistors all have the same value, and so if one resistor gives a time of 10 minutes, two produce a time of 20 minutes, three give 30 minutes, and so on. The timing period can be varied from about 2,700RC to 4,100RC seconds using R7, and in practice this is adjusted by trial and error to give a suitably high level of time accuracy.

The ZN1034E has an internal 5V shunt regulator, and R8 is the load resistor for this. The trigger input (pin 1) is connected to the negative supply so that circuit is triggered at switch on. The output at pin 2 goes low during the timing period, cutting off both Q1 and Q2. These two devices, which form the basis of an audio oscillator driving LS1, are biased into operation at the end of the timing period when pin 2 goes high, causing the alarm to sound.

The circuit consumes about 8mA under quiescent conditions, rising to about 16mA when the alarm is operating. The times provided by the unit can be varied to suit individual requirements but the timing resistance should be no more than about 10 megohms, and the timing capacitance should not be less than 3n3.

Timer



An adjustable timer, giving a delay of about 10 seconds to 1 minute, can be used for photographic and other purposes, or with various games where each competitor must make his move within the agreed period.

When the switch is moved to the "On" position, timing begins and C1 commences to charge through R1 and RV1. The two resistors R4 and R5 hold the source of Q1 at a fixed potential. When the voltage across C1 has reached a high enough level, Q1 gate is positive so that drain current flows through R3. This causes a voltage drop in R3, so that the base of Q2 moves negative. Q2 conducts and collector current flows in the relay coil, closing the relay contacts. When the switch is returned to the "Off" position, C1 is discharged through R2, so that the interval can be repeated.

A 2N3819 is suggested for Q1, and AC128 for Q2. With C1 as shown (470uF) the interval was found to lie between 10 seconds with a total of 250k in the R1/RV1 position, up to 1 minute with 2 megohm. Increasing C1, R1 or RV1 will lengthen the interval. Smaller values here will reduce it. This was with current rising to 40mA, with a 100 ohm relay.

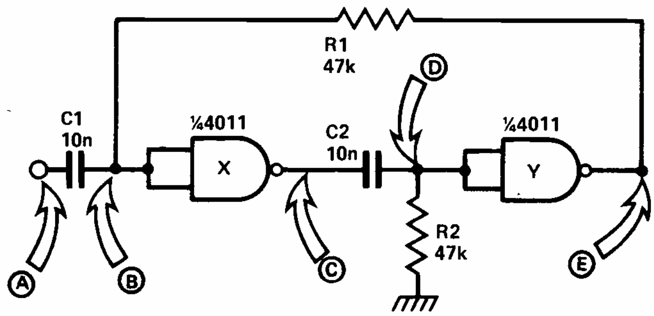
It is not of course essential that these values or transistor types be followed exactly, and other relays would also be practicable, provided Q2 allows a satisfactory current and voltage to suit the relay. Generally, a relay with a coil resistance of about 100 to 250 ohms will be most satisfactory.

The relay contacts can be wired so that when the relay coil is energised, the circuit is completed, or interrupted. For games and similar purposes, a 12 volt 3 watt indicator lamp can be operated from the same 12V supply. Should any kind of mains-voltage circuit be controlled, the relay must be a type intended for this purpose, and care must be taken to arrange mains circuit so that no danger can arise for the user.

CMOS Monostable

MONOSTABLE PERIOD
T = 0.69(R1 x C1 + R2 x C2)
IF R1 = R2 AND C1 = C2
THEN T = 1.38(R1 x C1)

IF R = 47k, C = 10n
THEN T = 0.7ms



A monostable is an electronic circuit that generates an output pulse of defined duration when triggered by an input signal transition. The monostable action can be used for many different functions. For example, when a cheap push button is used to trigger digital devices, spurious pulses often find their way into the circuitry. This is due to the contracts inside the button bouncing when it is pressed and released. Digital circuitry will regard all of the bounces as valid input signals and act accordingly – this can be disastrous in counting applications. A monostable can be used to de-bounce the push button; by setting the output pulse duration for a period longer than the longest expected bounce time, the bounces will have no effect on the main circuitry. So, with a simple monostable between each push button and the input circuitry, the digital devices are protected from the horrors of untamed push buttons.

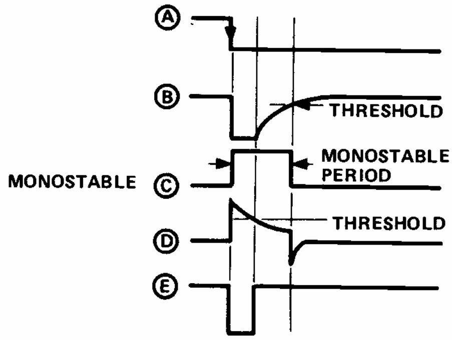
Only two inverters are needed for a monostable circuit. Here we have used a 4011B quad NAND package, as they are even cheaper than a CMOS inverter package and work just as well.

The monostable period is initiated by the transition of point A from positive to zero volts. This is called a negative-going edge. Transition of point 'A' in the opposite direction (a positive-going edge) will have no effect.

A negative edge at 'A' causes point 'B' to go to zero volts momentarily. This will then drive point 'E' to zero and this will hold point 'B' at zero volts, even when the pulse at 'A' is finished. The circuit will stay in this state while C2 charges. When C2 is fully charged there will no longer be a current through R2 and point 'D' will fall to zero again. 'E' will go positive and so will 'B', after a time determined by the values of C1 and R1. Point 'C' will fall to zero and C2 will be discharged, ready for the next pulse from 'A'.

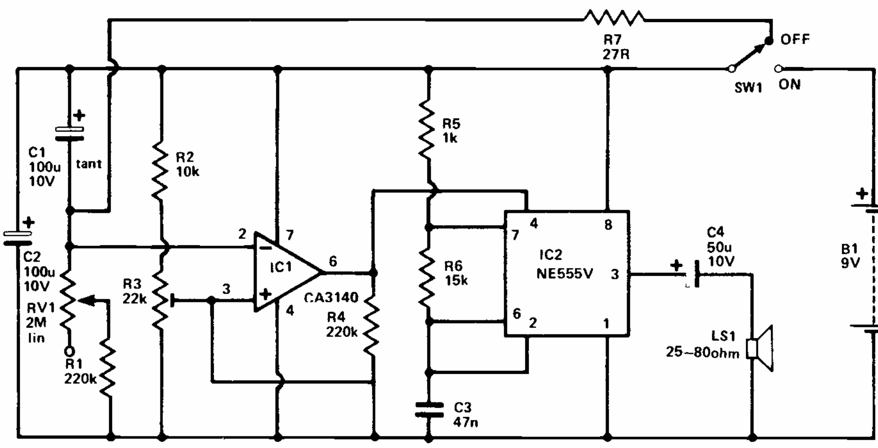
The length of time for which 'C' remains positive is determined by the values of C1, R1, C2 and R2. If either the resistance or the capacitance is increased, the time for which the monostable is triggered will increase.

MONOSTABLE WAVEFORMS



The equations for calculating the monostable period are given on the circuit diagram Polarised capacitors should not be used, and the resistance value of R should be kept within the range of 10kR to 10MR.

Simple Timer



This general purpose timer gives an audible alarm some predetermined time after the unit is switched on. With the specified values the time is variable from about 30 seconds to 5 minutes, but this can be altered to suit individual requirements.

When the unit is switched on, C1 begins to charge via R1 and RV1. Initially the voltage at the inverting input of IC1 will be higher than that appearing at the non-inverting input, and so IC1 output will assume a very low voltage. As C1 charges up, the voltage fed to the inverting input gradually falls until it starts to go below the voltage at the non-inverting input. IC1 output then begins to rise and due to coupling through R4 this increases the voltage at the non-inverting input. This causes a further increase in output voltage, and a regenerative action takes place which causes IC1 output to rapidly swing to almost a full positive supply potential.

The 555 is used to generate the alarm signal. The charge rate of C1 and thus the length of the timing interval can be altered by changing the resistance of RV1. The time delay is approx. 1.4RC (with C in uF, R in Meg., and the time in seconds), but due to the high tolerances of the timing components it is impossible to obtain highly predictable results. R3 has therefore been included so that the trigger voltage of the circuit can be varied, and by the trial and error R3 can be adjusted to give the appropriate timing range.

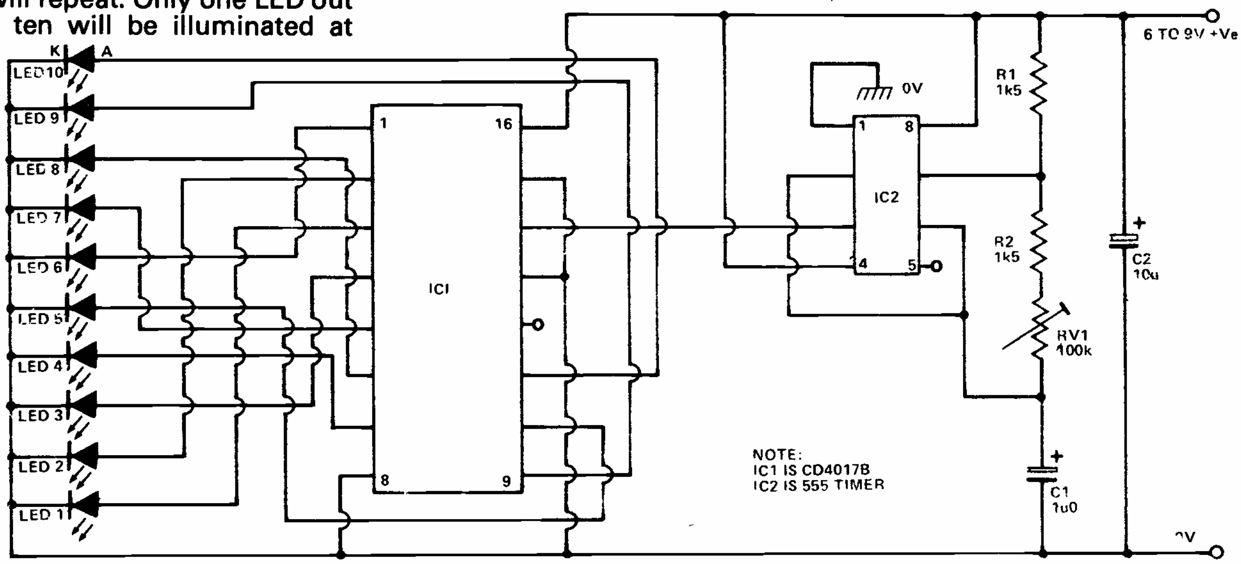
SW1 discharges C1 through current limiting resistor R7 so that the unit is ready to start a new timing run almost immediately.

LED Chaser

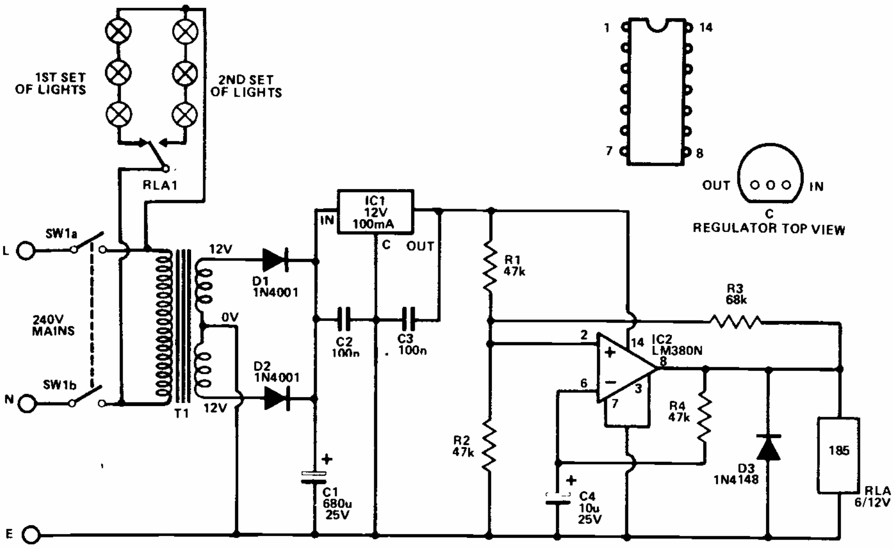
This unit produces a very attractive visual display in the form of a moving 'dot' on a column of ten LEDs. If the unit is wired up exactly as described in following paragraphs, the display will be such that an illuminated dot will appear to move sequentially and smoothly along the line of LEDs, from the bottom of the column to the top, until the top-most LED is reached, at which point the sequence will repeat. Only one LED out of the ten will be illuminated at any given moment of time.

The display can, if preferred, be wired up in a random fashion, so that the dot appears to jump about on a column of ten LEDs, but does so on a continuously repeating pattern. In either case, an attractive display will be produced. The unit can be used to simulate miniature shop display signs on model railway layouts, etc.

IC1 is a CD4017B decade counter with ten decoded outputs, and is 'clocked' by an astable multivibrator formed by IC2 and it's associated components. The action of the circuit is such that nine out of the ten decoded outputs of IC1 are low at any given moment of time, the remaining output being high. Each time that a clock pulse arrives a different output switches high, but all outputs go high in a fixed sequence. Each output is fed directly to it's own LED, so that the LEDs also switch in a fixed sequence. The clock rate is variable over a wide range via RV1.



Christmas Tree Lights Flasher



The usual method of getting the lights on a Christmas tree to flash on and off is to use a bimetal-strip type flashing bulb in the series chain of bulbs. As this switches on and off it breaks the circuit to all the bulbs so that they switch in unison. One drawback of this system is that most flashing bulbs provide a rather irregular flash rate; another is that it cannot be used to operate two sets of lamps.

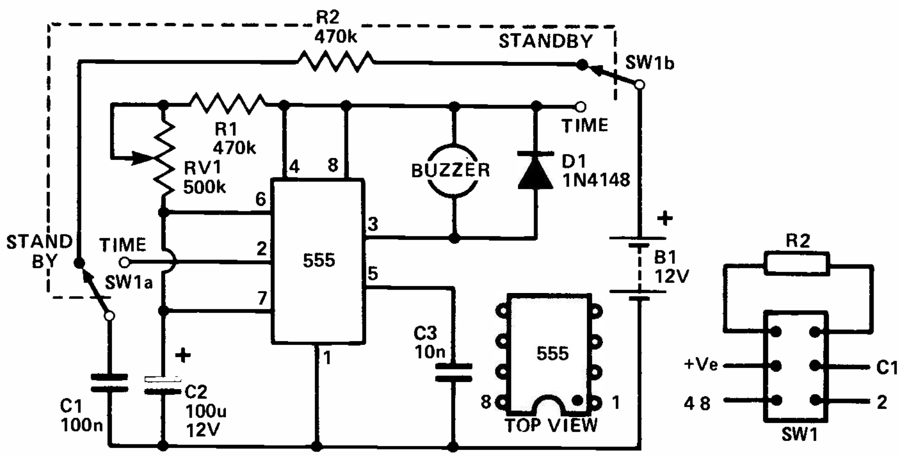
Both these problems can be overcome by using the simple circuit shown here. It is a low frequency oscillator (about 0.5 Hz) which controls the lights via a relay. Thus the lights are switched on for periods of about one second in duration at intervals of roughly one second. By using a changeover relay contact it is possible to use two sets of lights with the relay switching the power alternately from one set of lights to the other. If this alternate mode of operation is not required, then one set of lights is simply omitted.

The unit is powered from a simple stablised mains power supply having on/off switch SW1, stepdown and isolation transformer T1, push-pull rectifier D1 and D2, smoothing capacitor C1, and 12V monolithic regulator chip IC1. C2 and C3 aid the stability and transient response of IC1, and should be mounted physically close to this component.

A well known oscillator configuration is used here, but it is a little unusual in that it employs an audio power amplifier IC, rather than the more normal operational amplifier device. However, the LM380N audio IC can be used in operational amplifier type circuits. In this application it has the advantage of having a power output stage that can directly drive the relay with the high current it requires. D3 is used to suppress the high back EMF which is generated across the relay coil as it de-energises, and which could otherwise destroy IC2.

The switch on and switch off times of the circuit are proportional to the value of C4, and if desired they can be altered by changing the value of this component.

Egg Timer



The 555 (IC) is one of the most useful timer devices available. In this circuit it is wired up as a 'triggered monostable.'

Capacitor C1 is normally charged via R2 when the unit is in standby mode; when power is applied (switch SW1 to time), the trigger pin receives a pulse from C1, and the output goes high for a period determined by C2, RV1 and R1. When the time period is over the output pin returns to a low voltage and the buzzer turns on, indicating – your egg is ready! D1 is to prevent reverse induced voltage caused by the buzzer damaging the 555. When switched back to standby the buzzer supply is removed and C1 is recharged. The time period can be varied between 2 and 3 minutes with the component values shown. For shorter periods reduce C2 or R1; longer periods are given by increasing C2. Values of R1 and RV1 in total should not be more than 1MR, and C2 should not be more than about 500u – larger values will probably result in erratic time periods. C2 should have low leakage current characteristics if long time periods are required. If erratic time periods are produced a pushbutton (normally open-circuit) reset switch can be connected across C2. This will ensure complete discharge of C2 between cycles, if operated prior to each timing period.

The buzzer can be any 12V type, as long as it consumes less than 200mA, as this is the maximum the 555 can provide. Alternatively, a relay can be used to operate a remote buzzer or light.

Table Lamp

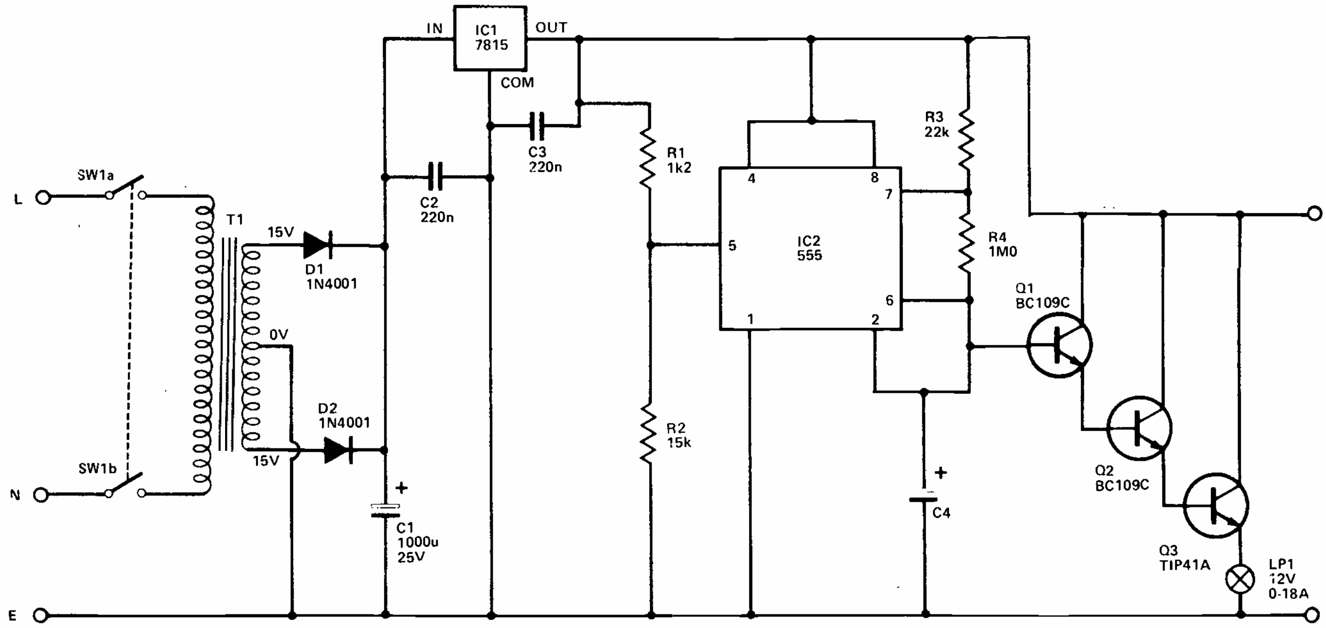
This circuit is for a decorative table lamp that produces a slowly changing background colour. The unit can have a number of lightbulbs of different colour, each modulated at a different rate, giving constantly changing blends of colour.

The circuit is powered from the mains via a stabilised power supply. T1 provides safety isolation and a voltage step down; its centre tapped secondary feeds a straightforward push-pull smoothing and rectifier network. The resultant DC supply is fed to regulator device IC1, which gives a stabilised 15 volt output. C2 and C3 are de-coupling components and should be mounted physically close to IC1.

The remaining circuitry is used to control one lamp. This must be duplicated for each additional lamp, and for best results at least three lamps should be used. There is no need to duplicate the power supply which can power up to five lamp circuits.

IC2 is a 555 device used in the astable mode. In this case R1 and R2 have been used to raise the two threshold voltages from their normal levels in order to make the unit more efficient. The roughly sawtooth waveform produced across C4 is used to control the lamp by way of a triple Darlington buffer stage.

The value for C4 gives a cycle time of about 4 seconds per uF. A three lamp version could therefore have values of (say) 10uF 33uF, and 100uF. These must be good quality components, preferably tantalum bead types.



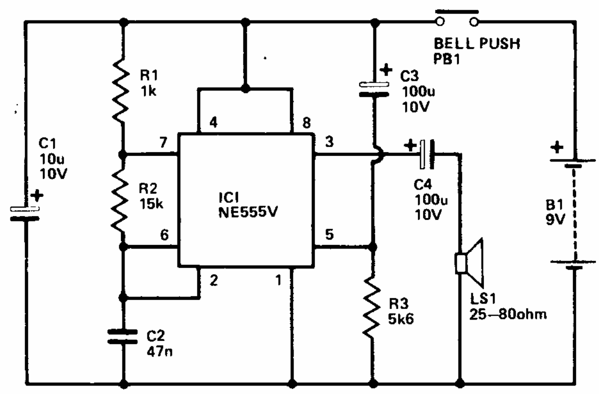
Doorbuzzer (1)

This design provides a novel doorbuzzer signal which starts as a low pitch and gradually rises in frequency.

The normal method of oscillation for the 555 is for the timing capacitor (C2) to charge up two thirds of V+ via two timing resistors (R1 and R2). The IC is then triggered and C2 is discharged through R2 and an internal transistor. The IC resets when the charge voltage drops to one third of V+, with the discharge transistor switching off and C2 commencing to charge up to the trigger potential once again.

This particular circuit does not oscillate in precisely this way, since the network comprised of R3 and C3 is used to shunt the potential divider (within the IC) which sets the trigger voltage. When SW1 is initially closed, C3 will be discharged and the trigger voltage will be raised. This increases the charge and discharge times of C2, and reduces the frequency of operation. C3 is quickly charged through R3 though, and after about one or two seconds the trigger voltage will have fallen to a level set by R3 and the internal potential divider. R3 pulls the trigger voltage below its normal level, reducing the charge and discharge times of C2 and causing an increase in the operating frequency. Thus, as C3 charges up, the output frequency is swept upwards, producing a novel and effective signal.

The main output at pin 3 of the 555 goes high during the charge period, and low during the discharge period, producing a rectangular waveform of low enough impedance to drive a speaker with up to a few hundred milliwatts of signal.

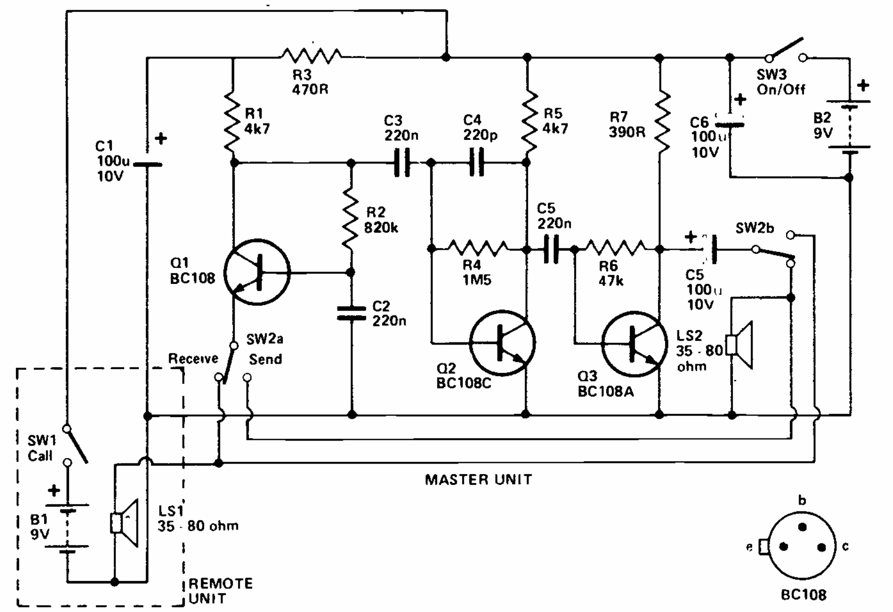


House and Home

Home Intercom

This intercom uses a straightforward three transistor amplifier which gives quite a good quality output (by intercom standards) and an adequate output power of a few tens of milliwatts.

As is normal practice with intercom designs, the loudspeaker in each station also doubles as a moving coil microphone. The position of SW2 determines whether the slave unit is 'sending' and the master station is receiving, or vice versa. Ideally this should be a biased switch which automatically returns to the 'receive' position when released. If it is not a biased switch, it could be left in the 'send' mode, preventing the remote station from calling the main one. SW3 is the ordinary on/off switch at the master station.



The amplifier is a three stage unit with capacitive coupling between stages. A common base input stage (Q1) is used as this gives a low input impedance. This is desirable as it minimises stray pickup of mains hum and radio interference in the connecting cable, and it also gives a good match to the microphones. The following two stages are both straightforward common emitter amplifiers. C4 rolls off the high frequency response of the circuit and this aids stability. It can also help to prevent RF breakthrough.

The prototype was tried with connecting cables up to about 10 metres or so long, and gave perfectly good results. It should work with considerably longer connecting cables if necessary. A three core connecting cable is required and this can conveniently be thin, three cored mains lead.

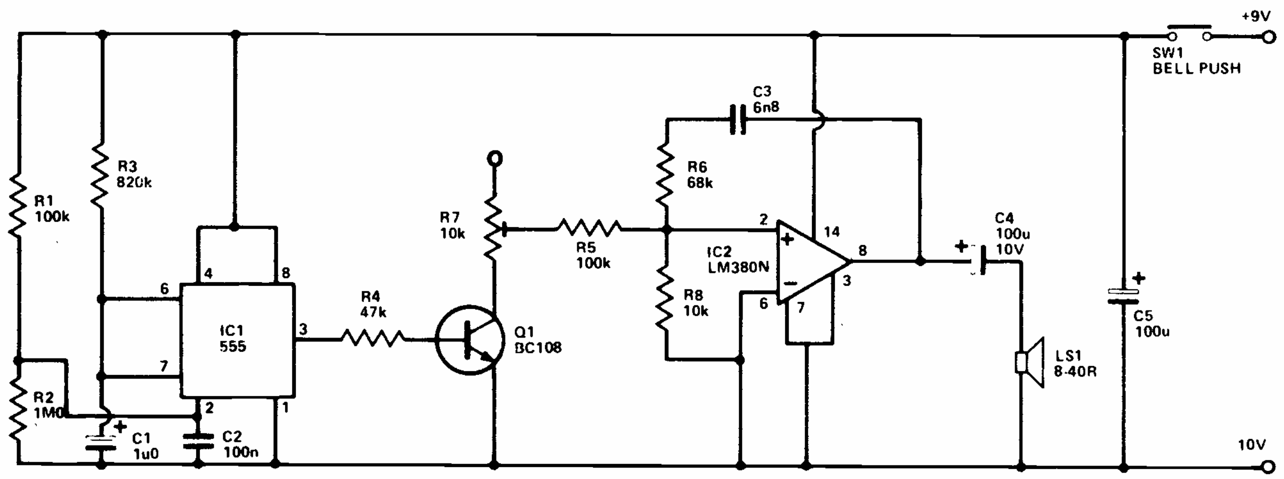
Doorbuzzer (2)

This is a two-tone doorbuzzer of the type that produces an initial tone for about one second, followed by a tone of lower pitch.

The audio tone is generated by IC2 which is an LM380N audio power amplifier. The frequency of oscillation is governed by the values of R6, R8, and C3. The specified values give an operating frequency of about 500Hz. The output from IC2 is fed to a loudspeaker via DC blocking capacitor C4, and an output power of nearly 1 watt rms is obtained using an 8 ohm speaker.

The two-tone effect is obtained by the inclusion of IC1 and its associated components. This is a 555 IC connected in the monostable mode; it produces a positive output pulse of just under one second in duration (set by R3 and C1) when a negative trigger pulse is applied to its pin 2. Such a pulse is produced at switch on, since C2 will initially be uncharged and will take pin 2 of IC1 to the negative supply potential. C2 rapidly charges by way of R1 though, so that the trigger input is quickly taken positive, and does not remain negative at the end of the output pulse (this would have the effect of lengthening the output pulse). When power is removed from the circuit, R2 rapidly discharges C2.

Q1 is biased hard into conduction by the output pulse from IC1, and it therefore effectively connects the series resistance of R5 and R7 in parallel with R8. This increases the operating frequency of the oscillator by an amount that is controlled by R7. At the end of the pulse from IC1, the oscillator operates at its nomal, lower frequency, giving the required two-tone effect.



Power Supplies

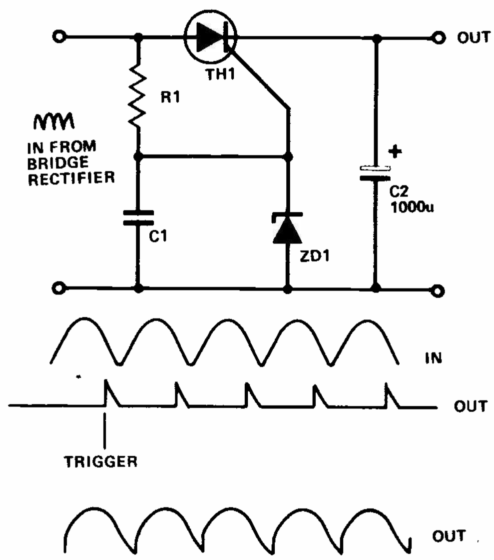
Simple Voltage Regulator

This is a voltage adjuster which enables you to obtain several switched values of DC voltage from a transformer without using tappings. It's particularly useful for lead-acid (car or motorbike) battery charging, as the voltage is so easy to change. The principle is that the zener diode, which may be one of several selected by a switch, applies a steady voltage, equal to the zener diode voltage, to the gate of the thyristor. At the anode of the thyristor, the waveform is a full-wave rectified voltage from the bridge rectifier. The cathode of the thyristor is connected to the reservoir capacitor C2, a large value electrolytic rated at the full peak voltage of the transformer secondary.

To see what happens, imagine that the peak voltage of the rectified wave is 25V and that we have a 15V zener diode. When the circuit is first switched on, capacitor C2 is uncharged, so that the cathode voltage of TH1 is zero. When the voltage at the anode starts to rise the thristor will not conduct right away, because C1 has to be charged up first. As C1 charges, however, TH1 will switch on, so that current flows, charging C2 to the full peak voltage.

Now if there is no load resistor connected across C2, the thyristor will not conduct again even if it is triggered, because the anode voltage will not be any higher than the cathode voltage. Any normal power supply is used with a load, however, so that we can assume that the voltage across C2 will drop quite a bit between the time of the first voltage peak and the next one. If the voltage across C2 is higher than the zener diode voltage though, the thyristor will not trigger, and the next voltage pulse does not cause the thyristor to conduct.

This continues until the output voltage drops to less than the voltage of the zener diode. When this happens, the thyristor will start to conduct whenever the anode voltage is higher than the cathode voltage, and the reservoir capacitor C2 will once again be charged to the full peak voltage. The thyristor will conduct every now and again; enough to keep the **average** voltage at the output



fairly steady at around the zener diode voltage.

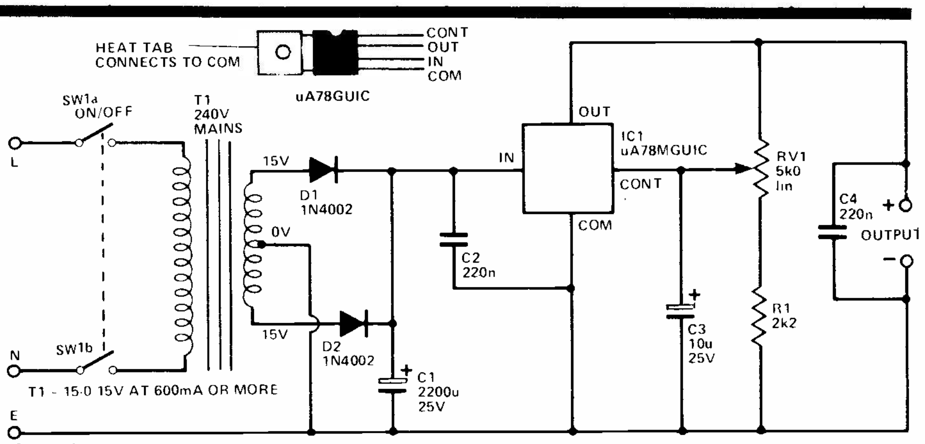
If this circuit is used as a battery charger (lead-acid only) then C2 can be ommitted.

To change output voltage, just switch in another zener diode! The output is not easy to smooth to an acceptable standard for most electronics equipment, but is ideal for battery charging or running lights or motors.

Single IC Power Supply

The unit described here has an output voltage variable from 5 to 15 volts, and a maximum output current of 500mA. The output is extremely well stabilised and noise is well below 1mV.

The mains supply is connected to the primary winding of isolation and step-down transformer T1 through on/off switch SW1. The centre tapped secondary of T1 feeds a standing fullwave rectifier and smoothing circuit which uses D1, D2 and C1.



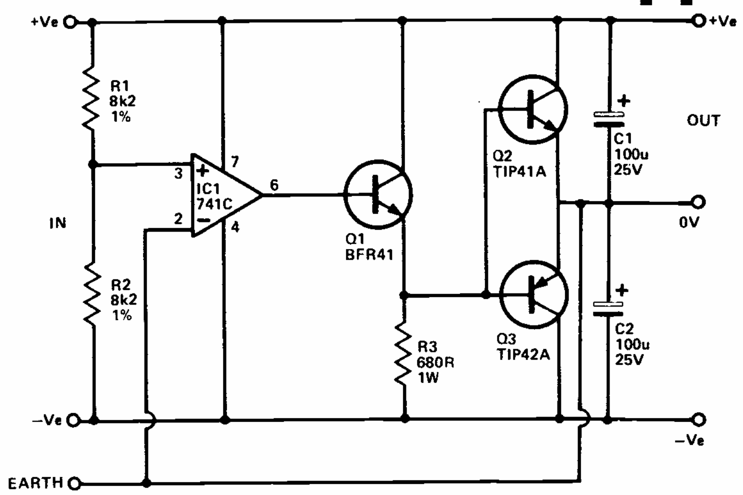
IC1 is the voltage regulator chip, and has four terminals. The unregulated input voltage is applied to the "IN" and "COM" terminals, while the stabilised output is taken from the "OUT" and "COM" terminals. The fourth terminal is "CONT" (Control) and if this is fed from the output, via a potential divider, negative feedback action will stabilise the voltage at this terminal at a nominal level of 5 volts. In this case the potential divider is formed by RV1 and R1. If RV1 slider is at the top of its track the output will be stabilised at 5 volts. A higher voltage would take the "CONT" terminal (which is now directly connected to the output) above 5 volts, causing the error to be sensed and corrected. Similarly, a lower voltage would take the "CONT terminal below 5 volts, causing the output to swing more positive.

If RV1 slider is moved down its track, the voltage fed to the "CONT" terminal will decrease, sending the output higher in order to return this potential to 5 volts. Thus RV1 can be used to vary the output voltage, with a maximum potential of about 16 volts or so appearing at the output when RV1 slider is at the bottom of its track. Above this level, at high output currents, there will be insufficient input voltage from the rectifier and smoothing circuit to properly maintain the output voltage.

The regulator device has built-in foldback current limiting which prevents the output from much exceeding 500 mA in the event of a minor overload. Stronger overloads result in decreased output current, the short circuit current being about 200mA!

Decoupling capacitors C2 and C3 should be mounted physically close to IC1, which must be mounted on a substantial heatsink.

Supply Splitter



Many operational amplifier based circuits require dual balanced power supplies, and cannot be powered direct from workshop units which have only a single output. An add-on unit, such as the circuit shown here, can be used to produce a low impedance centre tap on the output of a workshop power supply. If this centre tap is connected to the earth socket of the power supply, and made the 0V rail, the positive and negative rails will be at equal potentials relative to the 0V rail, but of opposite polarity. Thus the dual balanced positive and negative rails are provided, but each output rail is, of course, at a potential of only half the input voltage. In other words, for an output of (say) ± 15 volts it is necessary to feed 30 volts into the input.

This circuit will operate with input voltages of between 9 volts and an absolute maximum of 36 volts, and can handle output currents of up to about 500mA or so. The quiescent current consumption is 8 to 25mA depending on the input potential.

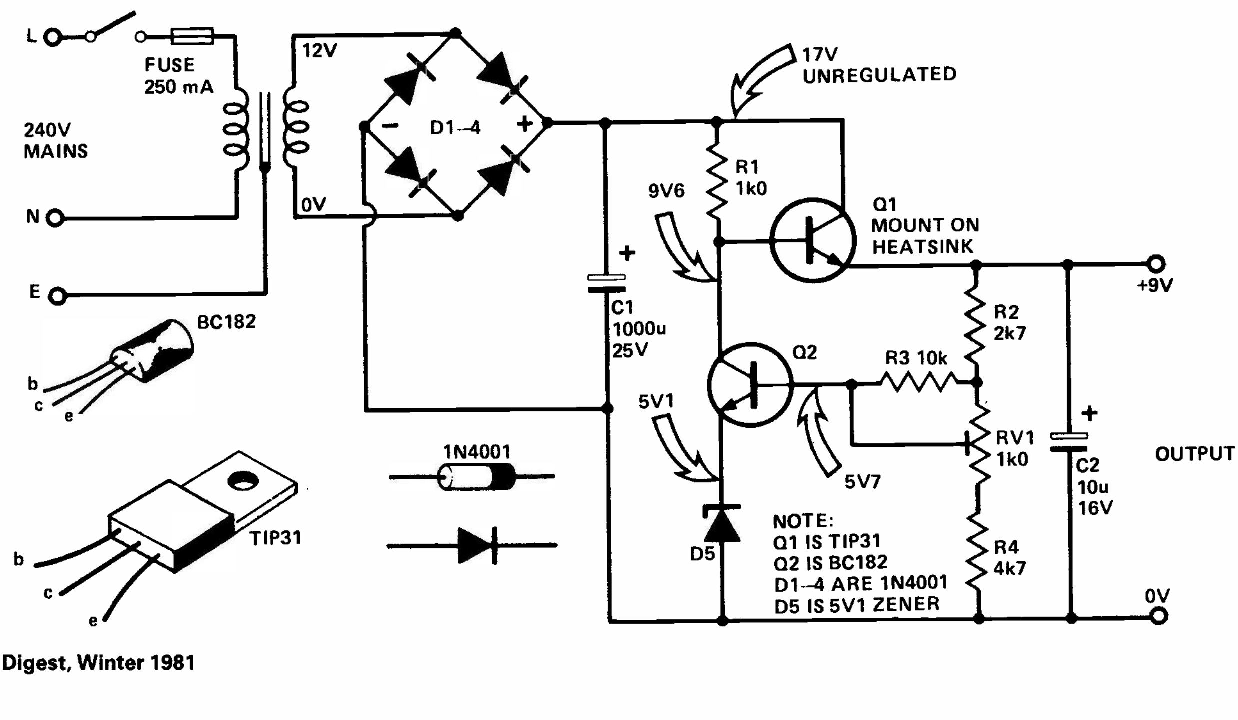
Basically the circuit consists of a potential divider circuit (R1 and R2) to produce an accurate potential of half the input voltage, and a buffer amplifier which is fed with this voltage and produces a low impedance output at the same potential. The buffer amplifier uses operational amplifier IC1 driving emitter follower stage Q1, which in turn drives complementary emitter follower stage Q2, Q3. There is 100% negative feedback from the output at Q2 and Q3 emitters to the inverting input of IC1 giving the amplifier unity voltage gain. The high gain of the circuit plus the negative feedback gives a low output impedance so that loading of the output does not pull the 0V rail significantly off-centre. A complementary output stage is used as this gives low quiescent current consumption.

If the unit is used at currents of more than about 100mA, Q2 and Q3 will probably need to be fitted with heatsinks to prevent them from overheating.

Regulated Power Supply

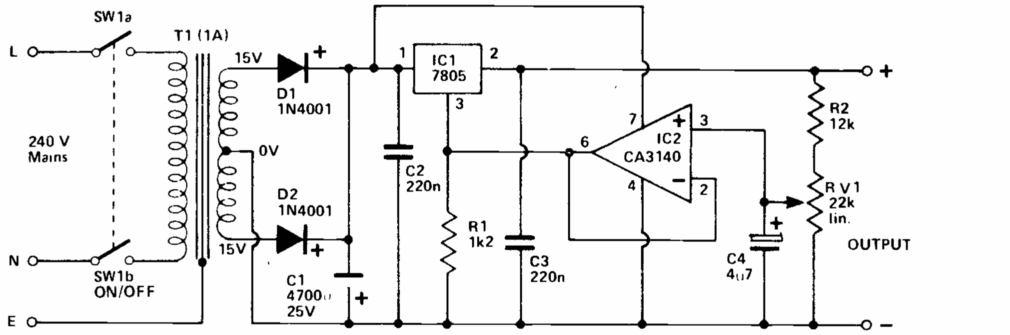
Two transistors and a voltage reference can be used to make a regulated power supply. Transistor Q1 is used as the power control element and so must be mounted on a heat sink. Q2 provides negative feedback and so helps iron out any changes at the output due to fluctuating load conditions or variations in the unregulated rail. The circuit operation is as follows: the current through Q2 and D5 sets up a voltage of 5V1 across D5. The base of Q2 is connected to the output by resistors R2, 3, 4 and RV1. If the output voltage rises, then more current will flow through Q2. This causes the voltage at the base of Q1 to fall, which in turn reduces the output.

RV1 is used to set up the output voltage to +9V. If the wiper of RV1 should accidentally lift off, the output voltage would instantly rise to that of the unregulated rail. To prevent this, R3 provides a permanent DC path to the base of Q2. C2 helps to improve the regulation when the load conditions change rapidly.



Power Supplies

5-13V Power Supply



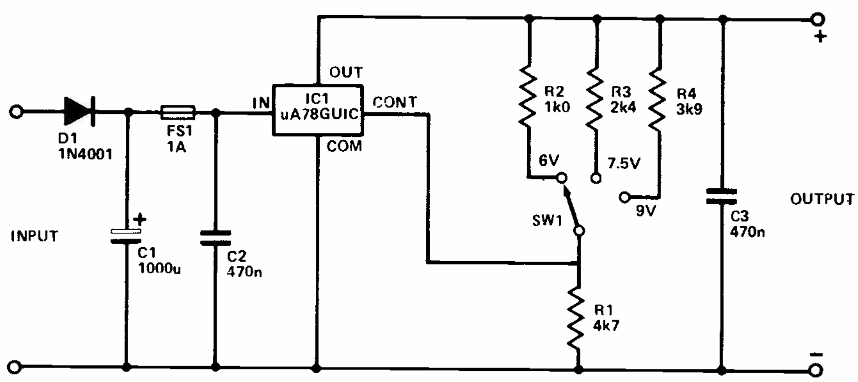
Although three terminal voltage regulators are often referred to as "fixed" voltage regulators, they can actually be used to provide output voltages other than their nominal ones, and can even be employed in variable voltage power supplies, as in the circuit shown here.

The three terminals of these voltage regulators are the input (pin 1), output (pin 2), and common (pin 3). The input voltage is applied to pins 1 and 3 with the correct polarity, and the stabilised output is extracted from pins 2 and 3. In effect, the device is actually stabilising the voltage on pin 2 at some fixed level above the potential at pin 3. Normally pin 3 is at 0V, and so the output voltage is determined by the nominal output voltage rating of the regulator.

Voltages greater than that for which the device is intended can be obtained simply by raising the pin 3 voltage by the appropriate amount. For example, a 5V regulator can be made to give a 9V output if its common terminal is taken to a potential of 4V. In this case a 5V regulator is used, and its common terminal is taken to a variable voltage of about 0 to 8V or so. This gives an output which is variable from about 5 to 13V. The voltage is supplied by R2 and RV1 which are connected across the stabilised output so that regulation efficiency is not significantly impaired. This voltage is at too high an impedance to directly drive pin 3 of IC1, and so a simple buffer amplifier based on IC2 is interposed between the two. R1 is a ballast resistor. A CA3140 device has been chosen for the IC2 position since the putput of this device can swing to within a few millivolts of the nagative supply voltage. Many alternatives such as the 741C device have a minimum output voltage of about 2V, which would give the power supply a minimum output voltage of approximately 7V, thus rendering the unit unsuitable for use with TTL and many other types of circuit.

The input voltage for the regulator circuit is derived from a conventional push-pull type step-down, rectifier, and smoothing circuit. C2 and C3 aid the stability of the circuit and should be mounted physically as close to the regulator IC as possible. C4 provides smoothing of the voltage at RV1 slider, and helps to give the circuit a low output noise level of about a millivolt or so. Regulation is also very good, the output falling by only 70mV between zero load and full output. The 7805 IC has current limiting circuitry which prevents an output current much in excess of 1A from flowing.

Car Cassette Power Supply



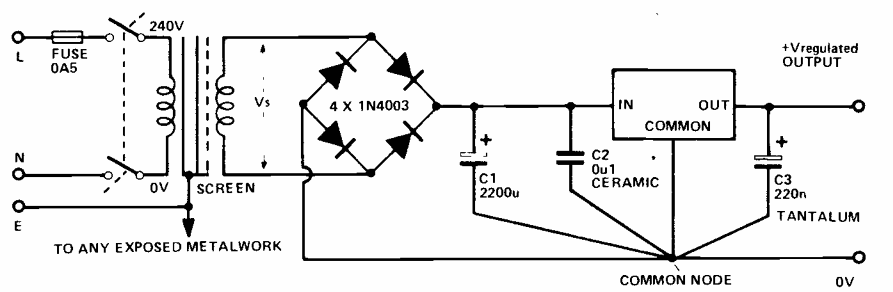
When using a portable cassette recorder in a car or boat it is far cheaper to run the unit from the vehicle's battery than from internal batteries. This circuit will give an output of 6, 7.5, or 9V from a nominal 12V input with a maximum output current of 1A.

It uses a four terminal monolithic voltage regulator to give the voltage stepdown and provide a well stabilised output. The input to the regulator is obtained via D1 which blocks the supply if it is connected with the wrong polarity, and fuse FS1. The input will probably contain a certain amount of noise which is smoothed by C1. C2 and C3 are decoupling capacitors which aid the stability of the circuit, and should be connected as close to IC1 as possible.

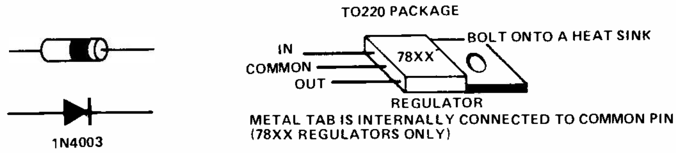
The output voltage of IC1 is set by a potential divider connected across the output and feeding into IC1's CONT (control) terminal. A negative feedback action stabilises the voltage on the CONT terminal at 5V, and the output at a level equal to 5V plus the potential dropped by the potential divider. The value of R1 sets the potential divider current at just over 1mA, giving a drop of about 1V across R2, 2V5 across R3, and 4V across R4. SW1 can thus be used to select output voltages of 6V, 7V5, and 9V by switching the appropriate resistor into circuit. If only a single output voltage is required, SW1 is omitted and the appropriate resistor is connected between IC1's CONT terminal and the output.

IC1 must be mounted on a heatsink, and this can be the case, if a metal type is used. The heattap of the device connects internally to the COM terminal, so the tab must be insulated from the case in the usual manner if the latter is earthed and the unit is installed in a positive earth vehicle. The uA78GU1C has thermal overload protection circuitry and foldback output current limiting incorporated in its design.

Power Supply With IC Regulator



REQUIRED OUTPUT VOLTAGE	REGULATOR 78XX	TRANS-FORMER SECONDARY Vs	CAPACITOR VOLTAGE C1	TYPICAL MAXIMUM INPUT VOLTAGE TO REGULATOR
5V	7805	8 TO 9 V RMS	16 TO 25 V	25V
12V	7812	12 TO 15 V RMS	25 TO 35 V	30V
15V	7815	15 TO 16 V RMS	35 TO 63 V	30V
24V	7824	20 V RMS	35 TO 63 V	38V



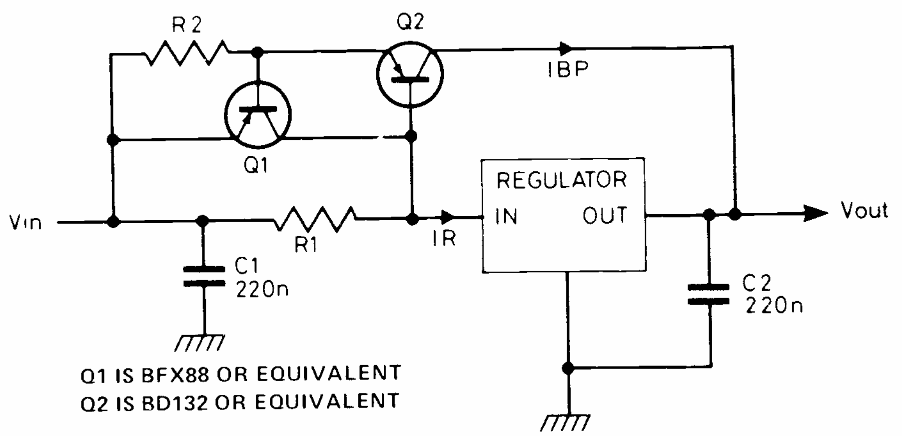
A regulated power supply used to require a fairly complex circuit but the introduction of special ICs has made matters much simpler.

The chart shows the various voltage requirements for four types of voltage regulator. Care must be taken that the voltage ratings of the regulator are not exceeded, as this can blow up the IC.

The secondary voltage from the transformer (it must be an isolating transformer but most are) is full-wave rectified by the diodes and then smoothed by capacitor C1. The voltage here is unregulated and has AC ripple superimposed upon it. The IC removes this and gives an almost ripple free, stable DC voltage. C2 and C3 must be sited close to the regulator to prevent any loss of performance due to high frequency instability. Note that the capacitor and common lead from the IC should be wired to the same point, this helps to reduce instability and hum problems that can occur due to poor layout.

The 78XX series of regulators which come in a TO220 package generally can supply 500mA. Therefore the maximum power dissipated in them is probably going to be 500mA times the voltage difference between the input and output terminals. This might mean that the regulator has to dissipate 5W of heat, therefore an adequate heatsink must be used. The 78XX series is current limited which means that if the maximum current is exceeded, the output voltage drops towards 0V. The regulator is thus short-circuit protected as long as proper heat sinking is provided.

Regulator Booster



If you've ever been in the unfortunate position of having a regulator in a piece of equipment which will supply, say, 0.5 A and you find that, for one reason or another, the equipment needs to draw 2A then this little circuit may save the day.

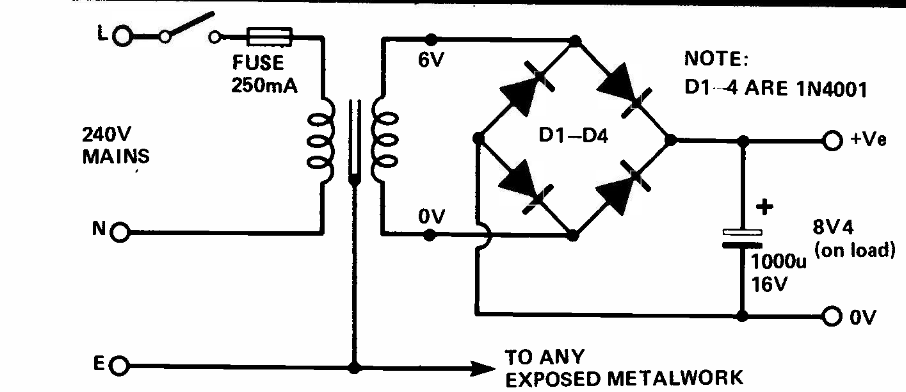
Essentially, it consists of a transistor (Q2) which will switch on at a certain current into the regulator and bypass it to an extent. It is similar to a 'current dumping' amplifier; the transistor will pass a large amount of current without regulating it and the regulator will do all the fine tuning to make sure the output is stable. The circuit also includes a second transistor, Q1, which will limit the current through the Q2 during a short circuit of the output; the regulator will presumably look after itself.

The operation of the circuit is fairly simple. As soon as the voltage drop across R1 is greater than the minimum bias voltage for Q2 (0V6), Q2 will turn on. As R1 is 1 ohm, this will occur when 0.6A is passing into the regulator. Having switched on, Q2 will supply current directly to the load. The regulator will sense the output voltage and if (as is likely, since the current through Q2 will be un-regulated) the output voltage is not now what it should be it will pass or hold back current as required to maintain the output voltage. The two capacitors are necessary to prevent the whole thing from going into a frantic oscillation as the regulator and Q2 try to decide who is doing what.

If the current through R2 rises above that needed to provide an 0V6 drop across it, Q1 will switch on and short out R1, thus removing the bias from Q2. This will prevent Q2 from passing more current than is good for it. Q2 should be mounted on a heat sink to prevent it from glowing.

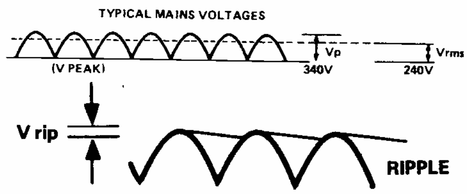
Power Supplies

Unregulated Power Supply



A single rail power supply is shown in the diagram above. For safety, the fuse is put in the live wire path to the transformer. Also, the live wire is connected to the 240V terminal of the transformer; this part of the primary winding is furthest away from the secondary and so increases the safety of the unit. The earth should be connected to any exposed metalwork and to the transformer screen, if it has one. The voltages quoted are AC voltages measured in Volts rms.

The output of the transformer is 6V rms on load. When the transformer is not loaded this voltage may increase by about 25%.



RIPPLE RULE OF THUMB

$$V_{rip} \simeq \frac{I_{load}}{C} \times (7 \times 10^{-3})$$

SO IF THE LOAD CURRENT (I load) IS 100mA

$$\text{THEN } V_{rip} \text{ IS } \frac{0.1 \times 7 \times 10^{-3}}{10^{-3}} = 700\text{mV}$$

12V DC Motor Speed Controller

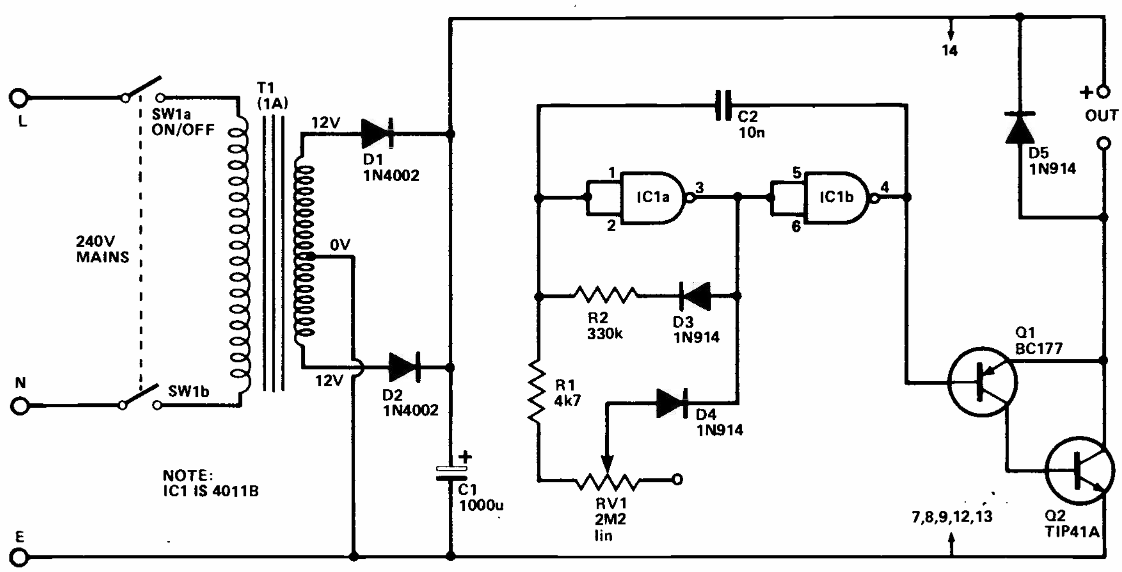
This motor speed controller is of the type where the motor is fed with a series of pulses of fixed duration and the power fed to the motor is varied by altering the frequency of the pulses. The higher the frequency, the greater the power fed to the motor, up to the point where there is virtually no gap between the pulses and the motor is operating at full speed. A useful characteristic of this type of controller is that it gives relatively good results at low speeds and wastes little power with consequent low dissipation in the circuit. This circuit is for 12V DC motors having a maximum current consumption of up to 1A (or 2A if the secondary current rating of T1 is raised accordingly.

T1, D1, D2 and C1 form a simple DC supply which gives a loaded voltage of just over 12V (just over 17V unloaded). This is used to drive a CMOS astable multivibrator which is based on two of the gates in a 4011B device. The two gates that are used each have their inputs connected together so that they act as inverters and are connected in what is basically the standard CMOS astable circuit.

The circuit differs from the standard configuration in that there are two timing resistances; one formed by R2 and the other by the series resistance of R1 plus RV1. D3 and D4 are steering diodes. In effect, R2 forms the timing resistance when the output of the astable is high and gives an output pulse of fixed duration. RV1 and R1 act as the timing resistance when the astable's output is low and the duration between output pulses can, therefore, be varied using RV1.

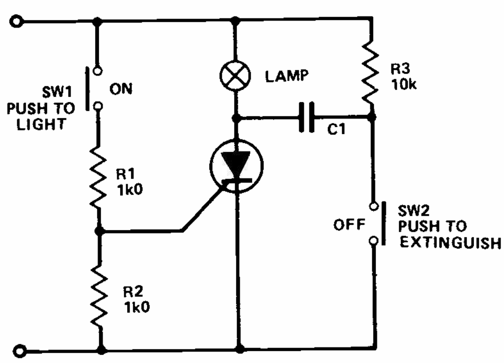
With RV1 at minimum there is a negligible gap between the output pulses, giving maximum speed from the motor. Increasing the resistance of RV1 increases the duration between the pulses, giving decreased average output power or no significant output power at all with RV1 at maximum resistance.

Since the astable has only a low output current capability, the motor must be driven via a buffer amplifier. This uses Q1 and Q2 in the common emitter mode, with 100% overall negative feedback so that unity voltage gain (but a high current gain) is obtained.



Thyristor Switching

In this circuit, the thyristor is triggered by the action of the push button switch – note the 1k resistor (R2) which prevents the gate from being triggered by stray radiated pulses. Once the thyristor has been triggered on there is a low-resistance conducting path between anode and cathode, so that the voltage across the thyristor is low, not much above the ninimum holding voltage at which the thyristor switches off.



The switch-off method uses a large-value capacitor C1 which charges up when the thyristor conducts. During the time that the thyristor conducts, the capacitor has one plate at supply voltage and the other at the low voltage of the thyristor anode. When the OFF switch is momentarily depressed, the plate of the capacitor which was at supply voltage is suddenly connected to zero volts. The other plate will follow it, dropping from about 0V2 to a negative voltage for a few milliseconds. This is time enough to allow the thyristor to become non-conducting. By the time the capacitor has re-charged, the thyristor is off.

This method works well when the load has a fairly high resistance (such as a low voltage lamp) and when a large value capacitance can be used. The capacitor shoud be a paper or plastic type, because electrolytics do not take kindly to having their plates at reverse polarity, which will happen if the OFF switch is kept closed while the capacitor re-charges through the load.

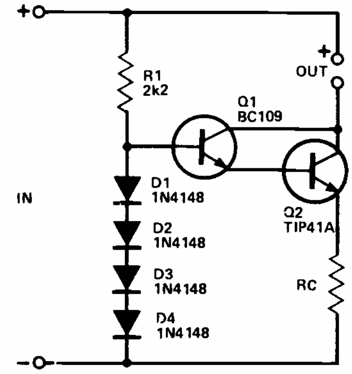
Power Control

Nicad Current Generator

This simple add-on circuit enables a DC bench power supply to be used as a Ni-Cad charger. These cells have a low internal resistance and can be damaged if the charge current is significantly higher than the figure recommended by the manufacturer. Furthermore, the cell voltage increases as charging progresses, making it necessary to steadily increase the charge voltage if the charge current is to be maintained.

This unit is a constant current generator circuit which limits the current fed to the Ni-Cad cell(s) to an acceptable level. In effect, the unit automatically adjusts the charge voltage to just the right level to give the desired charge current. The circuit is a standard constant current generator configuration with R1 and D1-4 being used as a sort of low voltage zener stabiliser. About 0V7 is developed across each of the four forward biased silicon diodes, giving a total zener voltage of about 2V8. Q1,2 are used as a Darlington pair and, therefore, have a very high combined gain so that quite high output currents can be produced by the fairly low drive current available. About 0V65 is dropped across the base-emitter terminals of both Q1 and Q2, giving about 1V5 across emitter resistor RC. The emitter current can be controlled by RC. The collector current of Q1,2 is virtually identical to the emitter current and is actually just fractionally lower as the emitter current is equal to the sum of the base and collector currents. Thus, provided a low impedance load (such as Ni-Cad cells) is present at the output, the current fed to the load can be set by giving RC the appropriate value.

The value of RC is equal to 1,500 divided by the required output current in milliamps and would, for example, be 10R for rapid charge Ni-Cads requiring a charge current of 150mA (1,500 divided by 150 = 10R).



The input voltage should be 3-6V more than the total voltage of the cells being charged. The cells should be connected in series across the output. Of course, the power supply must be capable of supplying the charge current drawn by the cells plus the additional few milliamps drawn by the current generator circuit itself. For charge currents of more than about 100mA it will probably be necessary to fit Q2 with a small finned heatsink to prevent it from overheating.

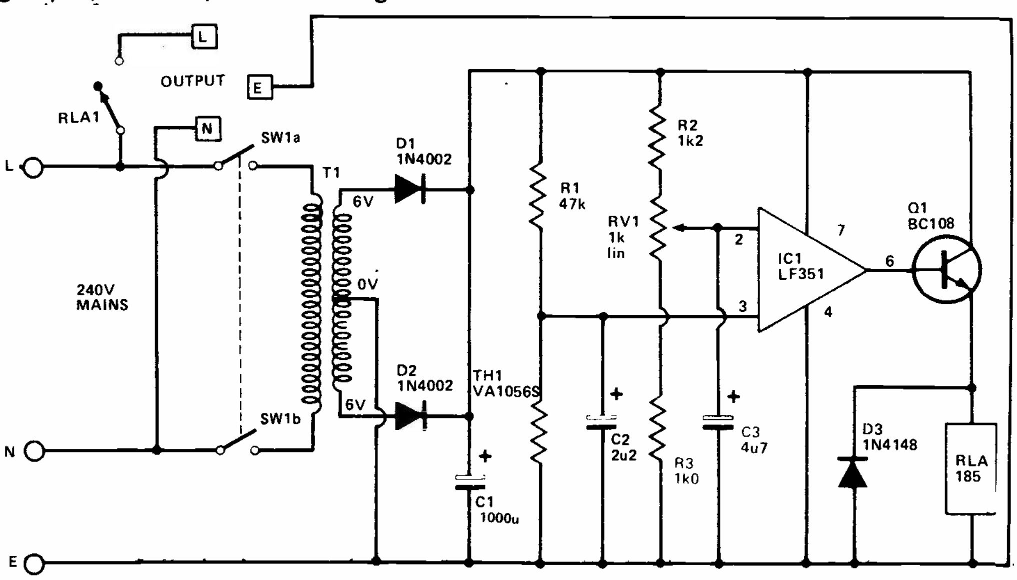
Thermostat

This simple thermostat will stabilise temperatures to within a fraction of one degree Centigrade. It covers a range which extends from a few degrees C to over 35 degrees C and is therefore suitable for use as a room thermostat, a thermostat for photographic solutions, etc.

The circuit uses operational amplifier IC1 as a voltage comparator. If the voltage at its non-inverting input is higher than that at the inverting input, the output goes to almost the full positive supply voltage. If the comparative input states are reversed, the output assumes a very low voltage.

The inverting input is fed with a reference voltage which is taken from the slider of RV1, acting as the temperature control. The non-inverting input is fed from a potential divider which consists of R1 and Th1.

Th1 is a negative temperature coefficient thermistor; its resistance varies according to temperature with increased temperature causing decreased resistance. If the voltage produced by Th1 and R1 is less than that set by RV1 slider, IC1's output will go high, switching on the relay via an emitter follower buffer stage, Q1. Power is then connected to the heater and the room temperature (or whatever) starts to rise. This causes a reduction in the resistance of Th1, and the voltage fed to IC1's non-inverting input. This voltage soon falls below that fed to the inverting input, causing IC1's output to go low, the relay to de-energise, and the heater to be switched off.



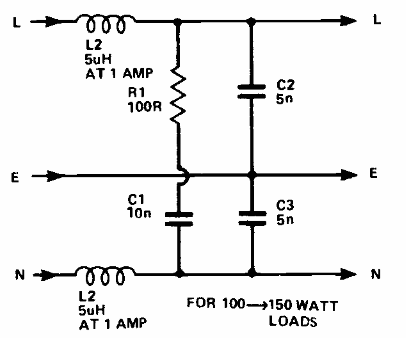
Mains Interference Suppression

If you were to take a look at the mains voltage on an oscilloscope you might expect to find a 250 Vrms sinewave. If you do see a pure sine wave then you are quite lucky and need not read any further! However, most mains supplies are composed of the sinusoidal generated by the electricity company *plus* various forms of interference generated by local users. When light bulbs, electric ovens and various electric motors are turned on they usually cause clicks and spikes to be generated. The worst offenders are large brush electric motors, thyristor dimmers and motor controls, spot welding units, in fact anything that grabs large chunks of current abruptly from the mains supply.

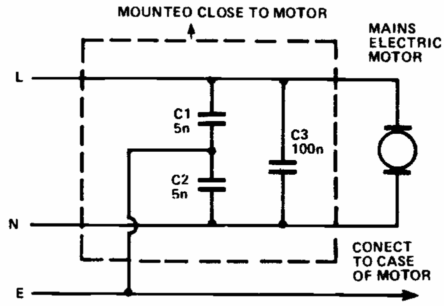
The clicks and spikes often make their way into electronic equipment, producing audible clicks on loudspeakers and generating errors in digital equipment. Faced with this problem there are two things that can be done: filter out the mains borne interference or try to prevent it being generated.

The top circuit shows a typical mains filter. The mains has to pass through a passive lowpass filter made up of an inductor and a capacitor, causing the high frequency parts of the interference to be attenuated. The inductor must have a high current rating (1 amp in this case), and the capacitors a high voltage rating (250Vac at least). This type of mains filter can be brought as a module for a few pounds, but it can also be made out of discrete components. In the latter case, use rubber sleeves on the connections.

The bottom circuit shows how to suppress motor generated interference and also reduce electromagnetic radiation. Again, the capacitors need to have a sufficiently large voltage rating, and take care to insulate all connections.

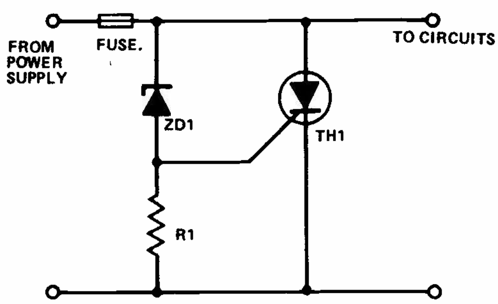


C1,2,3 MUST BE 250 VAC OR GREATER



Fast Fuse

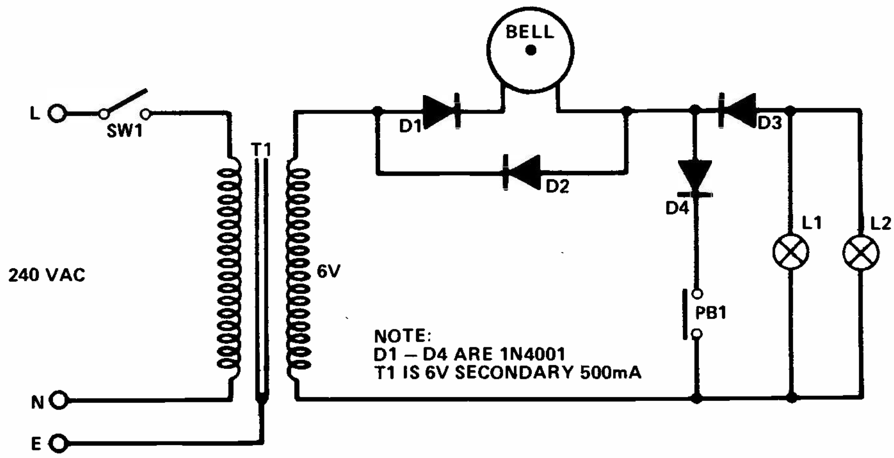
It may look a bit daft using a thyristor to blow a fuse, but it has two important advantages. The most obvious one is fuses don't blow just because the voltage rises, so that a fuse doesn't protect the circuit against excessive voltage, it only protects the power supply against excessive current. The other point is that a fuse takes some time to blow, several milliseconds. That may not sound like a long time, but several hundred pounds worth of IC's can be destroyed in only a thousandth of that time. The thyristor operates faster, getting the voltage down as soon as it reaches danger level, and blowing the fuse so that some attention is called for. The name "crowbar" circuit is a good one – the action is pretty much the same as that of putting a crowbar across the supply voltage!



In this circuit, Zener diode ZD1 has been selected so that it breaks down at 5V6. Now the normal voltage output of this circuit is 5V, and it's intended to supply a lot of IC's whose supply voltage must not rise much above 5V6. What happens if the voltage does get too high? Simple – ZD1 conducts and triggers the thyristor. The thyristor can then conduct, shorting out the power supply and causing the fuse to blow. Once the fuse has blown, there's no voltage in the circuit, and the thyristor instantly resets, ready to resume protection duty when the fault that caused the trouble is sorted out.

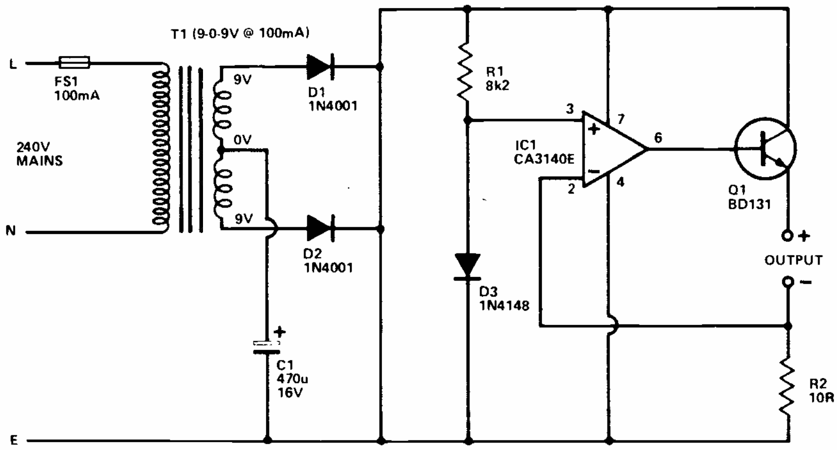
Two-Down-One

This circuit was designed so that people who have an illuminated door-bell can run a separate bulb over the door number without the need for a second set of wires from the transformer. If you try to use the same pair of wires to run a second bulb the increase in current will probably make the hammer of the bell 'tremble'. By using only one half of the AC cycle from the transformer to illuminate the bulbs whilst the other half of the cycle rings the bell, the problem is overcome.



Power Control

Nicad Charger

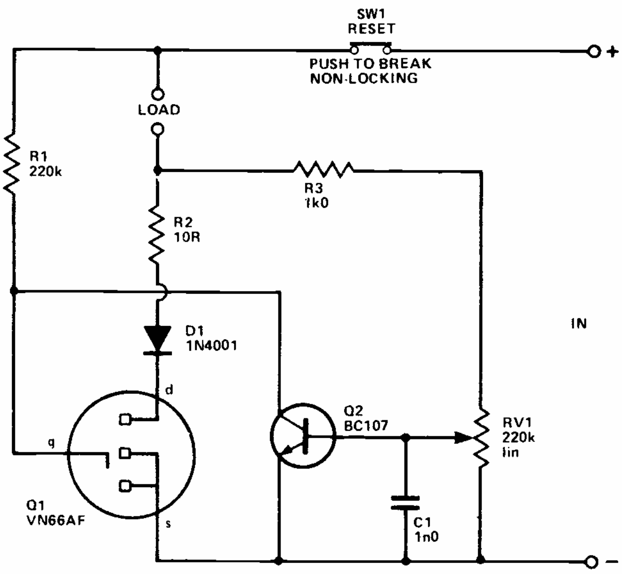


A special charger is needed for NiCad calls because they have a very low internal resistance, leading to an excessive charging current even if the applied voltage is only marginally too high.

In this circuit, T1, D1, D2, and C1 form a conventional stepdown, isolation, fullwave rectifier, and smoothing circuit. The other components provide the current regulation. IC1 is used as a comparator with discrete buffer stage Q1 giving a suitably high output current capability for this application. IC1's non-inverting input is fed with a 0V65 reference potential provided by R1 and D3. The inverting input is taken to earth by R2 under quiescent conditions, causing the output to go fully positive. With a NiCad cell connected across the output a high current will attempt to flow, causing the voltage across R2 to increase. It can rise to only 0V65. However, as a higher voltage reverses the comparative input levels to IC1, resulting in the output going lower and reducing the voltage across R2 towards 0V65. The maximum output current (the charge current) is therefore the current produced with 0V65 across 10 ohms, or 65 mA, in other words.

Some AA NiCad cells have a maximum recommended charge current of about 45 or 50 mA, and for these types R2 should be increased to 13 ohms in order to obtain the appropriate charge current. Some rapid charge types will take 150mA, and this requires reducing R2 to 4.3 ohms (3.3 ohms plus 1 ohm in series if a suitable component cannot be obtained). T1 should be changed to a type having a current rating of 250 mA, and Q1 should be fitted with a small bolt-on finned heatsink. The unit can charge up to four cells (six, if T1 is made a 12V type), and these must be connected in series across the output, not in parallel.

Overload Current Trip



This overload current trip can be used between the powered equipment and the power supply and will cut off the supply almost instantly if a preset threshold current is exceeded. The trip current can be varied from just a few hundred milliamps. The unit will work with supply voltages of 5-40V.

When power is first applied to the circuit, power FET Q1 will be biased hard into conduction by bias resistor R1. Power is, therefore, supplied to the load via Q1, D1 and R2. There will be a voltage drop across these components, and to some extent this varies with changes in the supply current. At low output currents there is likely to be a voltage drop of something in the region of 0V7, but this increases to a volt or so at high currents.

RV1 is adjusted so that at output currents below the required threshold level the proportion of the voltage dropped across Q1, D1 and R2 (and fed to Q2's base terminal) is not sufficient to switch on Q2. If the threshold current is exceeded, the voltage fed to Q2's base is then adequate to switch the device on and it diverts the bias current that formerly went to Q1's gate terminal. Q1 then switches off and cuts the supply to the load. Q2 remains switched on as it receives a strong base bias from the postive supply through the load, current limiting resistor R3 and RV1. Once tripped, the circuit thus latches in the "off" state. It can be returned to the "on" state by clearing the overload and then briefly operating SW1 so that the supply is momentarily disconnected from the unit. When the supply to restored it then starts at the "on" state once again. C1 ensures that the circuit always initially assumes the correct state and also helps to prevent spurious triggering of the unit.

When using the unit it should be kept in mind that about 1V is lost through the device and the output voltage from the supply must be adjusted to compensate. The current trip inevitably causes some loss of regulation efficiency, but this is only marginal. If the unit is to have a trip current of 100mA or more, R2 can be reduced to about 1R8 in order to maintain the low voltage drop and marginal degradation of regulation efficiency.

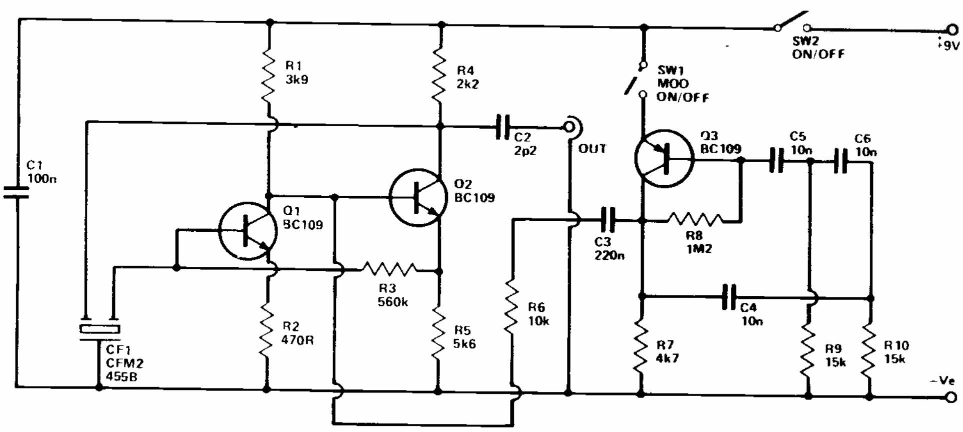
I.F. Alignment Oscillator

This simple piece of test equipment can be very helpful when aligning or realigning an AM superhet receiver. It provides an output at 455kHz, which can be modulated by an audio tone if the set is being adjusted for maximum AF output.

The circuit consists of two oscillator, Q1 and Q2 being used in the one which generates the 455kHz signal and Q3 in the one which provides the modulation signal. Q1,2 are connected in a straightforward two stage, direct coupled, common emitter configuration. However, neither of the emitter resistors are bypassed in this case as only a low voltage gain is required. The input and output are in phase and positive feedback between the two is provided by ceramic filter CF1 (available from Ambit International). A significant amount of feedback is only provided at the 455kHz operating frequency of the filter and so the circuit oscillates at this frequency.

A ceramic filter gives good frequency stability, requires no adjustment in order to produce the correct frequency and is cheaper than using a crystal. C2 provides DC blocking at the output, although it should normally only be necessary to connect the "hot" output to the receiver, no chassis connection being necessary.

A straightforward phase shift oscillator is used to provide the modulation signal and the specificed C-R values give an operating frequency of about 500Hz. C3 and R6 couple this signal to the base of Q2 where it amplitude modulates the 455kHz signal. SW7 can be used to cut the supply to Q3 and thus remove the modulation.



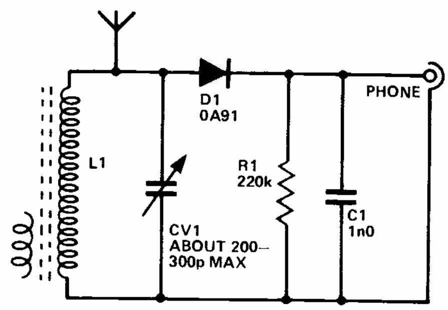
Crystal Set

The most simple form of radio for receiving broadcast stations is the crystal set, or more precisely, a modern equivalent using a semiconductor diode to provide detection. This simple set covers the normal medium wave broadcast band, has an output for a crystal earpiece and, in most areas, should give reception of Radios 1, 2 and 3 at reasonable volume (plus any local radio stations where these are in operation on the medium waveband). It requires no battery or other form of power source since energy derived from the received transmission is used to drive the earpiece. However, this does bring the disadvantage of needing an external longwire aerial to operate the set, as an ordinary ferrite aerial does not give sufficient pick up.

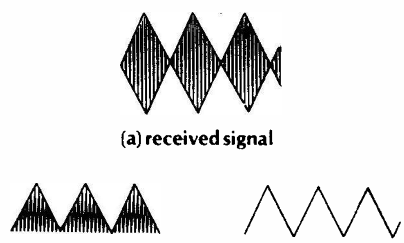
The tuned circuit is formed by L1 and CV1. This selects the desired transmission and rejects other stations. CV1 permits full coverage of the normal medium wave broadcast band to be achieved. In order to obtain good volume from the unit it is necessary to directly couple the aerial to the tuned circuit. For the same reason it is necessary to take the output to the detector direct from the tuned circuit. This inevitably gives the set rather poor selectivity, but it should still be adequate in this respect.

The form of modulation used on the medium wave band is AM (amplitude modulation). D1 half wave rectifies the RF signal to leave only the positive half cycles. R1 and C1 are used to smooth the RF half cycles, but their time constant is too short to produce a steady DC output. Instead the output rises and falls in sympathy with the mean RF signal level, so that the original audio signal is recovered at the output and fed to the earpiece.

The only adjustment the finished unit requires is to slide the aerial coil (L1) along the ferrite rod to find a position that permits full coverage of the medium wave band. The coil is then taped or glued in this position. The smaller winding of the ferrite aerial is not required and is either removed or just ignored. The aerial should preferably be an outdoor type about 10m or so long, but a few metres of hook-up wire fixed around the walls of a room or in a loft should give reasonable results.



NOTE:
L1 IS DENCO MW5FR FERRITE AERIAL OR SIMILAR.

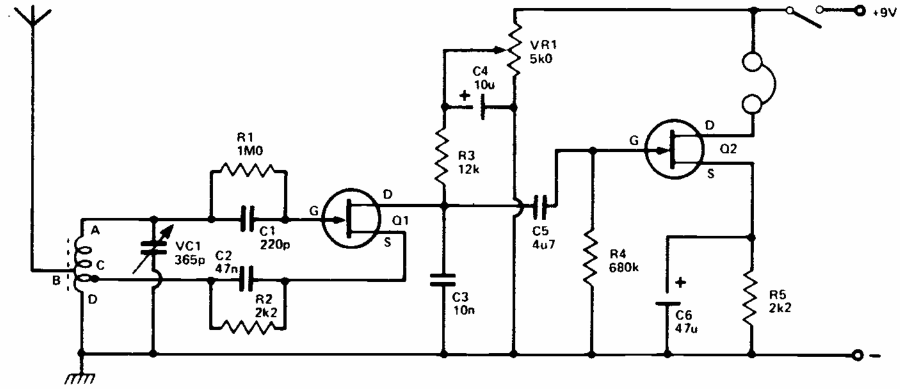


(a) received signal

(b) action of D1

(c) recovered audio signal

FET TRF Receiver



This circuit gives good headphone reception and can be constructed as a miniature receiver with a short throw-out aerial or it can be used with reduced range by relying on the ferrite rod alone.

Q1 is the detector and regeneration is obtained by a tap from the tuning coil to the source. The use of regeneration greatly improves selectivity and sensitivity to weak signals. Potentiometer RV1 allows manual adjustment of the drain potential of Q1 and so acts as a regeneration control.

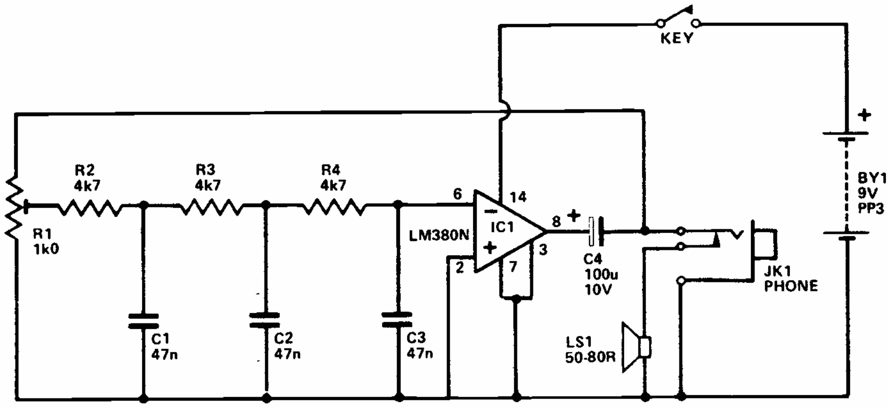
Audio output from Q1 is coupled to Q2 by C5. This FET is an audio amplifier, operating the headphones. A complete headset is preferable for general listening and phones of about 500 ohms DC resistance, or about 2k impedance, will give good results here. If a miniature earpiece is used, it should be a medium or high impedance magnetic unit. A crystal earpiece will require RC coupling.

The tuning inductor is fifty turns of 26swg wire, on a ferrite rod about 5in × 3/8in. If the turns are wound on a thin card sleeve which can be moved on the rod, this will allow adjustment of band coverage. The winding begins at A; aerial tapping B is at about twenty-five turns and D is the grounded end of the coil. The best position of the tapping C depends somewhat on the actual FET, on the battery voltage, and on whether or not the receiver is to be used with an external aerial wire. Should the tapping C be too near to end D no regeneration will be obtained or regeneration will be weak, even with RV1 rotated for maximum voltage. On the other hand, with too many turns between C and D, oscillation will begin with RV1 only slightly advanced. Best results are expected when regeneration begins smoothly with RV1 about halfway through its rotation. It was found that only one to two turns were required between C and D. The best method is to make C two turns from D then, if necessary, unwind half a turn or more at D.

When regeneration is obtained, a heterodyne will be heard if the receiver is tuned through a transmission. RV1 should then be turned back very slightly. Maximum sensitivity is achieved when Q1 is almost in oscillating condition. RV1 has to be set to suit the frequency tuned by VC1, so that final critical adjustment can be made. It is useless to regard RV1 as a gain control, and set it at maximum.

A metal case is suitable where an external aerial wire will be used. Where the ferrite rod only will be employed the box or case must be of plastic or other insulating material.

Morse Practice Oscillator



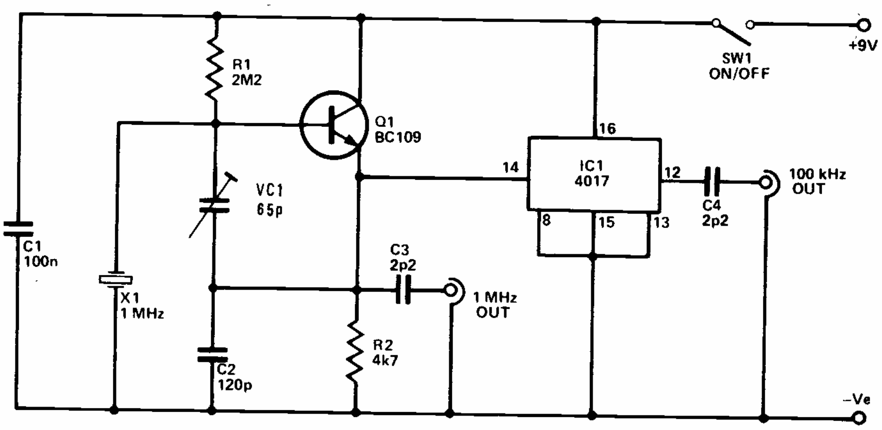
A Morse practice oscillator can be of considerable help when learning the Morse code. The simple unit descibed here has provision for an internal loudspeaker, and also has an output for a crystal earpiece, high impedance headphones, or a recorder.

The circuit is based on an audio power amplifier device (IC1) which is used in a phase shift oscillator circuit. Feedback is applied between the output and inverting (−) input of IC1 by way of a three section phase shift network. The three sections are formed by R2 − C1, R3 − C2, and R4 − C3, each of these sections providing 60 degrees of phase shift at a certain frequency. Thus, at this frequency there is a total phase shift of 180 degrees through the three sections. The circuit oscillates at approximately 1k5Hz with the specified values.

Ideally a circuit of this type should provide a sinewave output, as a pure tone is easy to listen to for long periods and is the waveform produced by an actual CW (Morse) transmission when it is resolved by a receiver. This circuit will provide a reasonably pure sinewave if the gain of the amplifier just slightly more than compensates for losses through the feedback circuit. This is achieved by adjusting R1 to give the appropriate loss level through the feedback path. In practice it is backed off close to the point where oscillation ceases, due to a lack of feedback.

The output signal is fed to the loudspeaker via C4 and a break contact on the output socket. The latter automatically cuts of the speaker when a plug is inserted into the socket. The unit provides an output power of about 100 mW rms, and has a current consumption of approximately 20mA with the key down.

Crystal Calibrator



A problem with home-constructed short wave receivers is that of providing an accurately calibrated tuning dial. A crystal calibrator solves this problem.

The calibrator circuit shown here has fundamental outputs at 100kHz and 1MHz. However, it does not merely provide calibration signals at these frequencies, but also at harmonics of these frequencies. The 1MHz output therefore provides calibration signals at 2MHz, 3MHz, 4MHz, etc., while the 100kHz output provides signals at 200kHz, 300kHz, 400kHz, etc. These additional frequencies are produced because the circuit is designed to give an output signal that is virtually a squarewave. This gives a signal rich in harmonics at frequencies that are readily detectable up to 30MHz (the upper limit of the short wave spectrum) on any reasonably sensitive receiver.

Q1 is used in a simple 1MHz crystal oscillator, and it operates in the emitter follower mode. VC1 and C2 effectively form a tap on the crystal which acts as a parallel tuned circuit. The output of Q1 is coupled into this tapping, and this gives the positive feedback path needed to produce oscillation. The circuit oscillates at the resonant frequency of the crystal since there is only an efficient feedback path at this frequency, sufficient to produce strong oscillation and an output rich in harmonics. A crystal is used rather than an ordinary L-C tuned circuit as a crystal gives better accuracy and stability. The 100kHz output is obtained by feeding the 1MHz signal to a CMOS 4017 divide by ten circuit.

VC1 must be adjusted to give optimum accuracy and this is easily achieved by connecting a short lead to the 100kHz output and placing it near a radio tuned to the BBC LW 200kHz transmission. This will produce a low frequency beat note (heard as a cyclic rise and fall in the volume of the station), and VC1 is simply adjusted for the lowest attainable beat rate. A beat rate of under one per second should be easily obtained.

10 To 30MHz Preselector

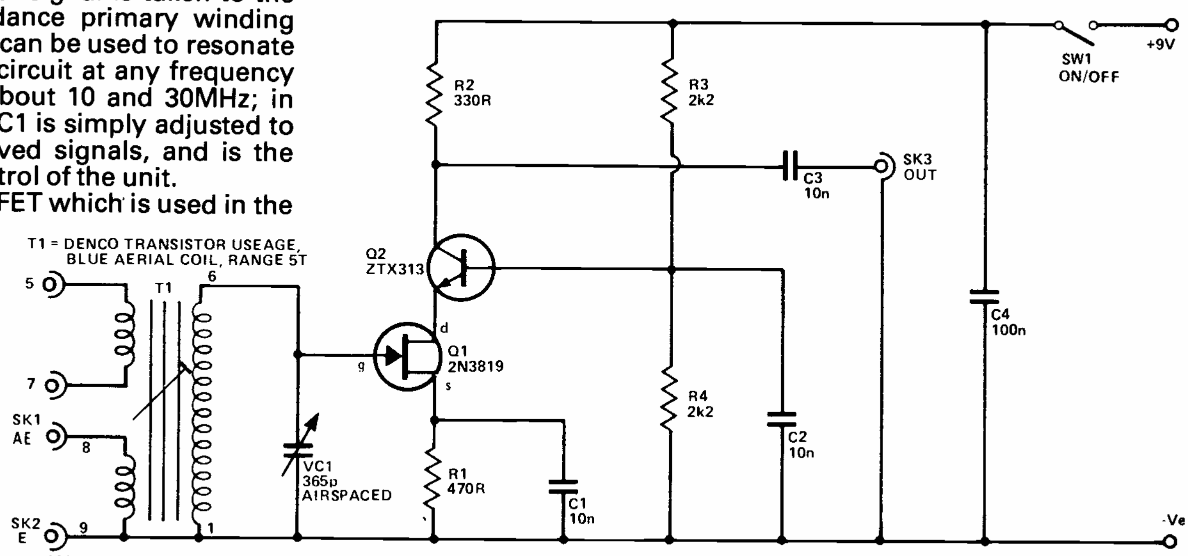
Many older or less expensive SW receivers give a relatively poor level of performance on the high frequency bands. One way of improving the performance is to add a preselector at the input.

A preselector is a tuned RF amplifier which boosts the aerial signal before it is fed to the receiver. Apart from giving improved sensitivity,the decreased RF bandwidth helps to attenuate any spurious responses by the receiver.

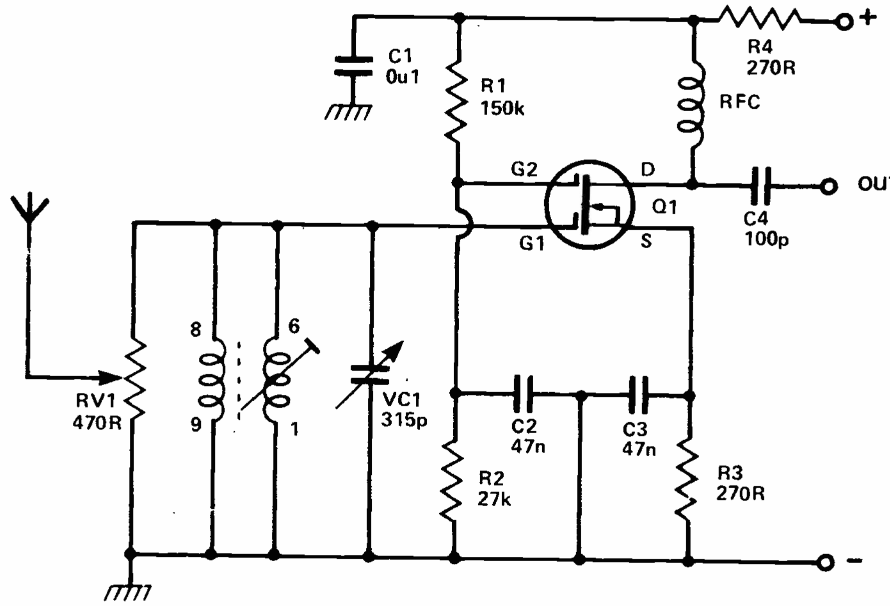
The aerial signal is taken to the low impedance primary winding of T1. VC1 can be used to resonate the tuned circuit at any frequency between about 10 and 30MHz; in practice, VC1 is simply adjusted to peak received signals, and is the tuning control of the unit.

Q1 is a JFET which is used in the common source mode, and has R1 and C1 as its source bias resistor and bypass capacitor respectively. It directly drives the input of Q2, which is an ordinary bipolar device used in the common base mode. This has R2 as its collector load, R3 and R4 to provide base biasing, and C2 as the base decoupling capacitor. This two stage amplifier is a form of "cascode" circuit, and gives good performance at the fairly high frequencies involved here. The voltage gain of the circuit is well over 20dB. C3 provides DC blocking at the output of the unit.

Construction of the unit is not critical, but try to keep all the wiring reasonably short. As supplied, the core of T1 is fully screwed into the former, and in order to obtain the correct frequency coverage the core must be unscrewed so that approximately 10mm of metal screwthread protrudes from the top of the coil. T1 can be mounted in a B9A valveholder, incidentally. The twin lead connecting the output of the preselector to the aerial and earth sockets of the receiver should be reasonably short to minimise losses.



Medium Frequency Amplifier



This circuit is primarily intended for use over the 1.7MHz to 30MHz range and will be found to provide considerable gain. RF amplifiers of this kind are generally used to improve long distance short wave reception, to increase volume, and to reduce second channel interference on the higher frequencies.

To avoid winding coils and permit easy band changing, Denco (Clacton) miniature plug-in coils may be used; these are the "Blue" (Aerial) ranges. The most useful coils will be Range 3, 1.67-5.3MHz, or 580 to 194 metres; Range 4, 5-15MHz, or 60 to 20 metres; and Range 5, 10.5-31.5MHz, or 28 to 9.5 metres. Exact coverage depends on the setting of the adjustable cores, and will also be modified if VC1 is of different value. The coils are inserted in a B9A type holder. If only a single range is wanted, the coil can be mounted by its threaded end, and leads are then soldered directly to the pins.

RV1 is an adjustable aerial input control, as overloading may easily arise with strong signals. R1 and R2 provide the voltage for gate 2, and R3 is for source bias.

The drain circuit is arranged for capacitive coupling by C4 to the aerial socket of the receiver. This lead should not be unnecessarily long, as this may cause losses, as well as picking up signals which cause second channel interference. If the lead is screened, it must be no longer than necessary. A 2.6mH short wave sectionalised radio frequency choke will be satisfactory for the frequencies mentioned.

Construction is best in a metal case, which can have a hinged lid if plug-in coils are to be fitted. No ganging difficulties can arise with VC1, which is adjusted for best volume.

Second channel interference is caused by signals which are 2×IF frequency from the wanted signals. With a 470kHz intermediate frequency, these offending signals will be 940kHz from the wanted transmission. As a result, interference from this cause is unlikely at low frequencies, but very probable at high frequencies. Such second channel interference is considerably reduced, or completely avoided, by using a tuned RF stage of this kind; actual results depend on the receiver IF, and frequencies tuned.

A 9 V supply is adequate and may be drawn from the receiver if convenient; only 2mA to 3mA or will be wanted. The MEM618, 40602, and 40673 will be found satisfactory here.

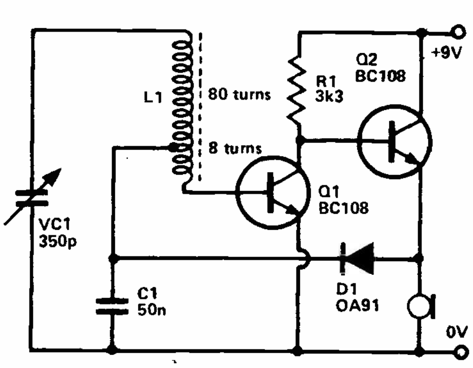
Two Transistor Radio

Although transistors are cheap, this was certainly not true some years ago, and transistors were made to work at maximum efficiency. A common technique in simple radio circuits was to use a reflex circuit; that is, one transistor amplified both at RF and AF.

The circuit shown uses very few components yet the operation is surprisngly complex. The coil L1 is made from 88 turns of enamelled copper wire, 32 swg for example, on a 5/16in ferrite rod about 4in long with a tap at 8 turns. The tuned circuit is made up from VC1, the 80 turns of L1 and C1. The latter has practically no effect upon the circuit as it is a very high value for RF purposes.

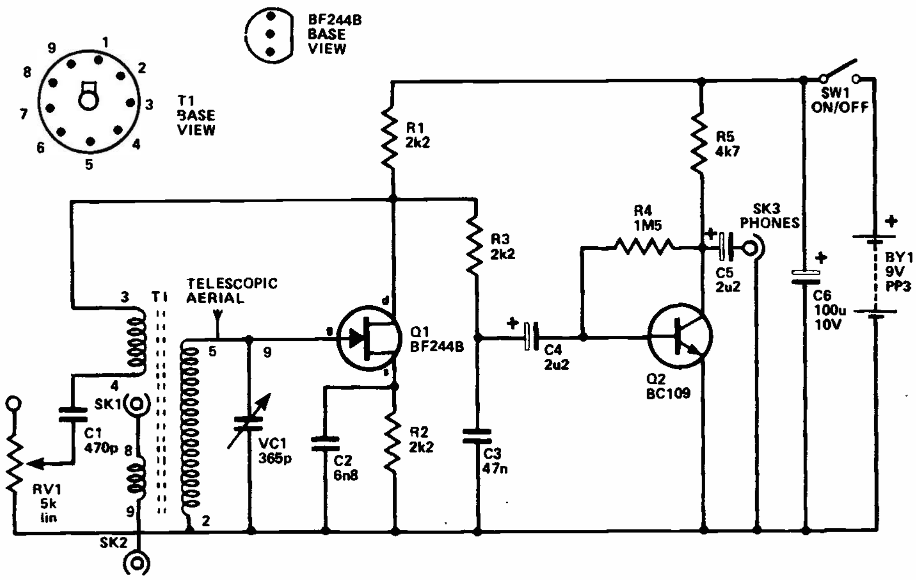
The RF signals picked up in circuit appear at a very high impedance which does not connect well to a regular transistor amplifier, but the 8 turns act as an autotransformer giving a good match to the base of Q1. This transistor amplifies the RF which is fed to Q2, acting as an emitter follower. The RF appears at the emitter of Q2 but the high impedance magnetic earpiece acts as an RF choke so the signal is passed through D1 and is detected by it and smoothed by C1. The signal is now at audio and it is connected to the base of Q1 via the 8 turns and is again amplified but this time it drives the earpiece; some audio is fed back but this acts as negative feedback. The base bias for Q1 is also supplied via the diode.

Most general purpose small signal transistors such as a BC 108 will work well in this circuit. R1 is worth experimenting with to obtain the best possible performance.



Note that most earpieces are 8 ohm types and will barely work in this circuit, high impedance types are less common than they used to be but are available – the impedance should not be less than 250 ohms. The value of VC1 is not very critical – almost all of the transistor type tuning capacitors will work.

Simple SW Radio



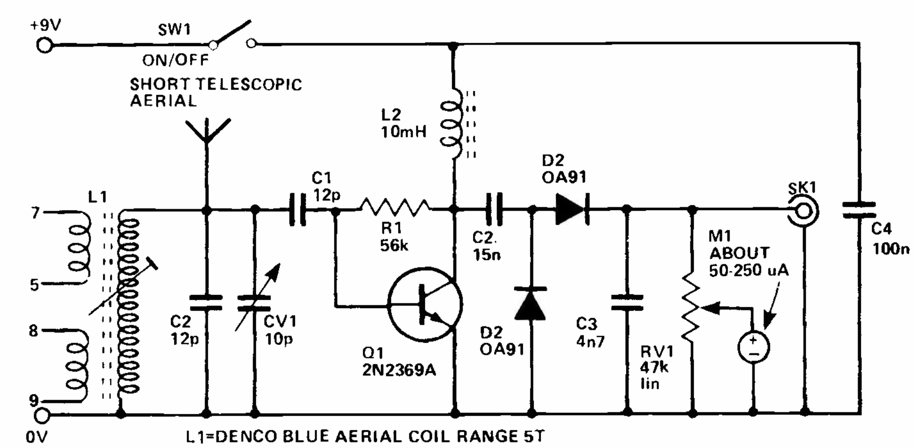
This simple SW set tunes from about 5 to 17MHz, covering the 19, 25, 31, 39 and 49 metre broadcast bands. It can be used with an external aerial and earth (which connect to SK1 and SK2 respectively), but it provides reception of a large number of stations using a short telescopic aerial.

VC1 is the tuning control, and forms the tuned circuit together with the main winding of T1. Signals from a long wire aerial are coupled to the tuned circuit via a small coupling winding on T1, but if a telescopic aerial is used it is necessary to directly couple it to the tuned circuit. Signals at the frequency selected by the tuned circuit are fed into the gate of Q1 which is used as a common source RF amplifier/regenerative detector. The process of regeneration merely entails sending some of the output from Q1 drain back to the input of the circuit so that it is amplified for a second time. This achieved by using a third winding on T1, with C1 providing DC blocking and RV1 controlling the amount of regeneration.

There are several reasons for using regeneration; the obvious advantage is that it gives increased gain and sensitivity. A less obvious one is that it increases the detection efficiency of the circuit. This type of detector relies upon the fact that it will amplify one set of half cycles more than the other set, giving a crude form of rectification. Thus, by adding an RF filter (R3 and C3) at the output of the amplifier, the RF signal is removed to leave the detected AF signal. Regeneration increases the inequality between the levels of amplification received by the two sets of half cycles, giving improved detection efficiency. Regeneration also produces improved selectivity, enabling the set to pick out just one station at a time from the crowded SW broadcast bands.

The audio output from the detector is considerably amplified by a common emitter stage based on Q2 before being fed to a crystal earphone or a pair of high impedance headphones. Current consumption of the set is about 2mA. Construction would not be difficult, but all the RF wiring should be kept reasonable short and direct. T1 can be mounted in a B9A valveholder by the built-in screwthread and nut.

27 MHz Radio Control Monitor



A radio control monitor such as that described here is invaluable when setting up a radio control transmitter. This circuit has good sensitivity and it is unnecessary to have the monitor in close proximity to the transmitter in order to obtain a signal of adequate strengths even when it is used with a low power transmitter.

The signal picked up by the telecopic aerial is coupled directly into the tuned circuit. The core of L1 is adjusted so that CV1 is able to tune the monitor to any frequency within the 27MHz band. The setting of this core is not too critical, since the unit covers somewhat more than the entire band.

Q1 is used as a common emitter amplifier with L2 as its collector load and R1 to provide base biasing. C1 couples the signal in the tuned circuit to the input of the amplifier. The value of C1 is chosen to give optimum signal transfer. The high impedance signal in the tuned circuit is matched to the low input impedance of Q1 by a sort of capacitive divider action (the input capacitance of Q1 forming the other section of the capacitive divider).

C2 couples the output of Q1 to a straightforward rectifier and smoothing circuit. This produces a positive voltage which is roughly proportional to the strength of the received signal. This voltage is used to drive M1, which gives a comparative indication of received signal strengths. An inexpensive meter is perfectly suitable for use in the unit due to the arbitrary scaling. RV1 can be used to reduce the sensitivity of the unit, if necessary. The modulation signal of the transmitter (if it is an AM type) can be monitored using a crystal earphone connected to SK1.

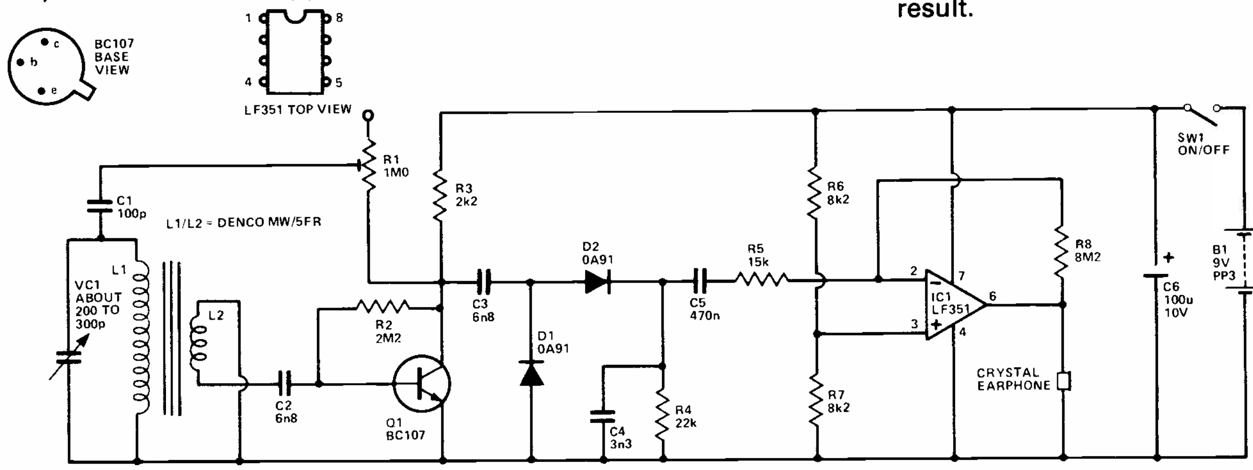
Simple M.W. Radio

This simple radio gives many hours of use from a PP3 battery, and will give good volume from a crystal earphone when tuned to all but the weakest of MW transmissions.

L1 is the tuned winding of the ferrite aerial, and VC1 is the tuning capacitor. The signal picked-up and selected by the aerial is coupled via low impedance winding L2 to a conventional common emitter amplifier stage based on Q1. Some of the output from Q1's collector is coupled back to the ferrite aerial by R1 and C1. R1 controls the amount of feedback, or regeneration as it is normally known in this application. One purpose of the regeneration is to produce increased amplification, and thus give improved sensitivity. It also improves the ability of the set to select just one of several closely spaced transmissions. This is very important, since the selectivity of the set is provided by a single tuned circuit, and performance would be very poor in this respect without the use of regeneration.

Most of the output from Q1 is fed to a simple diode detector circuit which uses D1, D2, C4 and R4. This halfwave rectifies the RF signal and smoothes the remaining RF half cycles to leave the audio signal. This is then fed to a low noise audio stage which uses IC1 in the inverting mode. IC1 is a low noise BIFET operational amplifier; its closed loop voltage gain is set by the ratio of R5 to R8 at about 550. The current consumption of the set is only about 2mA.

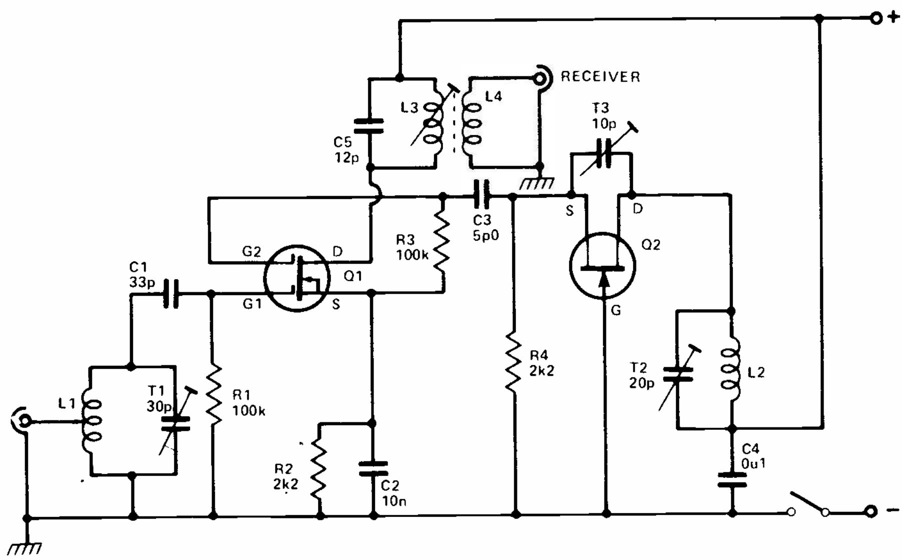
The aerial coil must be positioned on the rod so that the full MW band can be received. This is a matter of trial and error. The coil is then glued or taped in place. R1 is adjusted for the lowest value that does not cause the circuit to oscillate at any setting of VC1 (oscillation causes a tone to accompany the received station). If regeneration cannot be obtained, reverse the connections to L2. Keep the wiring to the "hot" ends of VC1 and L1 well away from the wiring to Q1 collector or uncontrollable regeneration may result.



144MHz Converter

The reception of 2 metre signals is generally possible using a converter and standard short wave receiver. With such an arrangement the 144MHz or other VHF signal is changed in frequency so that the converter output falls within the tuning range of the receiver.

A converter of this type often uses an RF amplifier and a low frequency crystal controlled oscillator followed by frequency multipliers. This allows high sensitivity and excellent frequency stability but is relatively complicated and expensive item. It is possible to use a much simpler circuit.



L1 is broadly tuned to the wanted frequency band by T1 and signal input is to gate 1 of Q1. Q2 is the local oscillator with the operating frequency determined by L2 and T2. Oscillator injection is via C3 to gate 2 of Q1. The frequency of the output from the drain of Q1 is the difference between G1 and G2 frequencies. Thus if the signal at G1 is 144MHz, and Q2 is tuned to oscillate at 116MHz, output will be at 28MHz. Therefore 144-146MHz can be covered by tuning the receiver from 28MHz to 30MHz. L3 is broadly tuned to this band, and L4 couples the signal to the receiver.

The oscillator can actually be tuned above or below the aerial circuit frequency of the converter, as it is the difference between converter signal input and oscillator frequencies which determines the converter output frequency. It is also possible to choose other reception and output frequencies, provided L1, L2 and L3 are chosen to suit.

L1 and L2 are wound in the same way, except that L1 is tapped one turn from its grounded end. Each coil has five turns of 18 swg wire, self supporting, formed by winding the turns on an object 7mm in diameter. Space turns so that each coil is ½in or about 12mm long.

L3 is fifteen turns of 26swg enamelled wire, side by side on a 7mm former with adjustable core. L4 is four turns, overwound on the earthed (positive line) end of L3. Layout should allow very short connections in the VHF circuits. A co-axial aerial socket is fitted near L1. A screened co-axial lead is preferred from L4 to the receiver to avoid unnecessary pick-up of signals in the 28-30MHz range. The converter will operate from 9-12V.

L3 should first be peaked at about 29MHz. If a signal generator is available, couple this to Q1 drain by placing the output lead near the drain circuit. Tune generator and receiver to 29MHz, and adjust the core of L3 for best results. Otherwise, couple an aerial by means of a small capacitor to the drain circuit, and tune in some signal in the 28-30MHz range.

It is now necessary to tune L1 to about 145MHz, and L2 to 116MHz, or 174MHz. If an absorption frequency indicator is available, this will permit an approximate setting of T2. A dip oscillator will also allow T1 to be adjusted. Subsequently adjust T2 to bring the wanted signals in at the required frequencies and peak these for best volume with T1, then check the setting of L3 core.

The converter is best assembled in a small aluminium box, completely closed, which can be placed behind the receiver. Note that if Q2 is not oscillating, no reception is possible through the converter. Q2 should be a VHF FET, such as the BF244, MPF102 or similar and if necessary T3 may be adjusted to secure oscillation. The 40602, 40673, or similar VHF types will be satisfactory for Q1. If needed, frequencies can be brought within the swing of T1 and T2 by stretching or compressing L1 or L2.

The aerial may be about 38½in long, constructed as a simple self-supporting or wire dipole, with a feeder descending to the converter. Amateur activity is most likely to be greater at weekends, and in many areas a whip or very short wire aerial will provide local reception.

Radio 4 Tuner

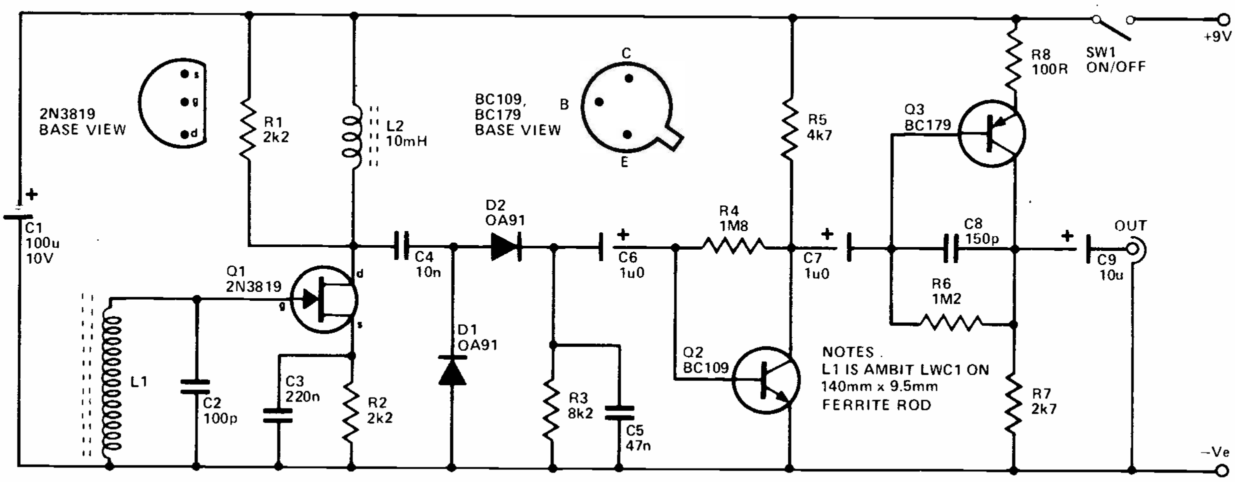
This tuner is designed for use with a tuner/amplifier which does not have long wave coverage and is therefore unable to receive BBC Radio 4.

However, it can also be used as a personal receiver for reception of Radio 4 if the output is fed to a crystal earphone or a pair of high impedance magnetic headphones.

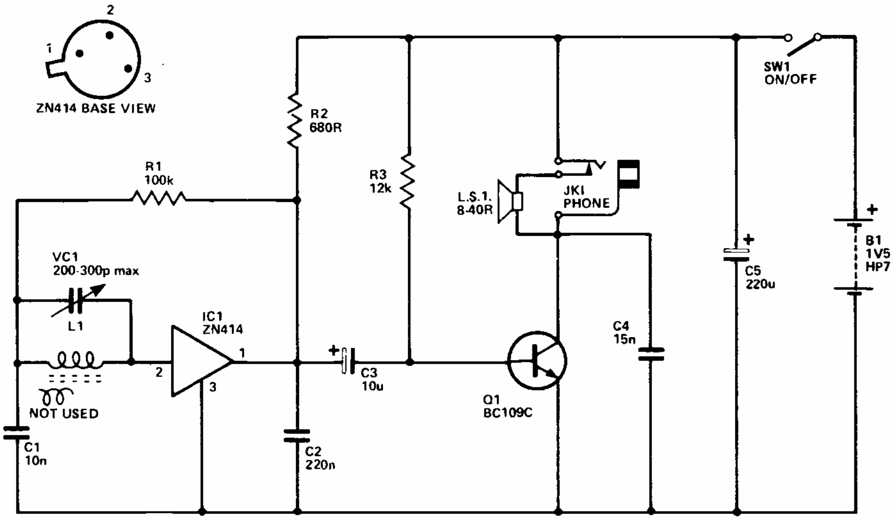
Q1 is used as a JFET common source amplifier and its gate terminal is fed direct from the ferrite aerial (L1). This is quite acceptable since a JFET has an extremely high input impedance and will not place a significant degree of loading on the aerial. L1 is used to bias the gate of Q1 to the negative supply rail. C2 brings the ferrite aerial to resonance at approximately the Radio 4 frequency of 200kHz, and L1 is simply slid along the ferite rod to tune the unit to the correct frequency. L2 forms the main source load for Q1, but it was found to be necessary to damp this using R1 in order to prevent instability.

The amplified RF output from Q1 is fed by C4 to a straight forward AM detector circuit, D1, D2, R3 and C5. The demodulated AF signal is then coupled by C6 to a high gain, low noise, common emitter amplifier based on Q2. This considerably boosts the signal, but it is still at an inadequate level to drive many amplifiers. A second common emitter stage using Q3 is therefore used to further boost the signal, and this gives an output amplitude of several hundred millivolts RMS. The full gain of Q3 is not required, and R6 is used to provide local negative feedback which produces the required reduction in gain. C8 rolls off the high frequency response; this aids stability and improves the signal to noise ratio of the unit.

The tuner only has one control, on/off switch SW1. The current consumption of the unit is about 2.5mA. When the correct position on the ferrite rod has been located, it should be firmly taped or glued in position.



Bedside Radio



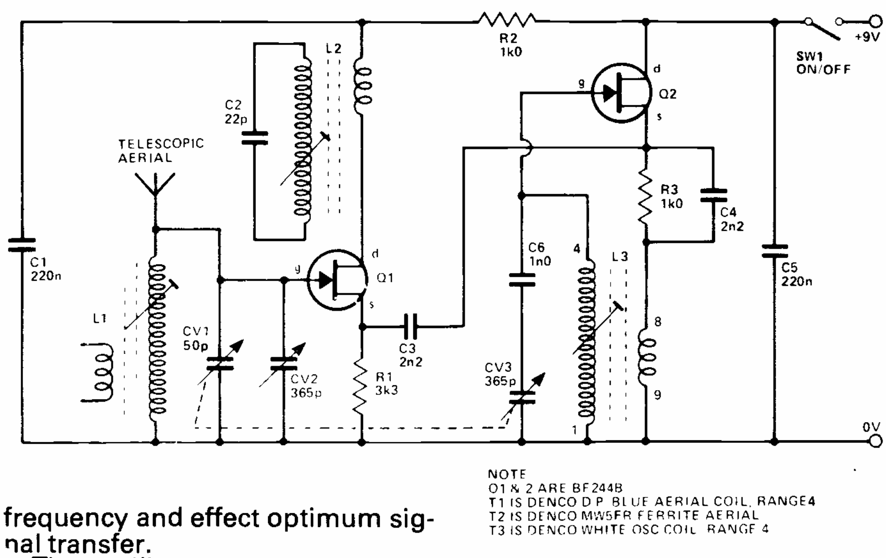
This very simple MW radio provides low but adequate loudspeaker volume for use as a bedside set, and also has provision for a low impedance magnetic earphone. The radio frequency circuitry is based on a ZN414 IC which provides considerable amplification for the signals picked up in ferrite aerial L1. VC1 is the tuning capacitor and R1 is a bias resistor for the ZN414. C1 provides an RF path to earth for the "cold" end of L1. The ZN414 has a transistor detector stage at the output, but this requires discrete load resistor R2 and RF filter capacitor C2 in order to give an audio output.

When strong signals are received there is a slight drop in the voltage at the output of IC1, effectively reducing the supply voltage fed to the device and giving a fall in gain. This produces a crude but reasonably effective form of automatic gain control (AGC) which reduces the possibility of overloading and gives a more consistent audio output level when tuning to stations of differing signal strengths.

The output stage is a simple common emitter class A type based on Q1. Interstage coupling is provided by C3, and R3 is the base bias resistor. The speaker or earphone form the collector load for Q1; JK1 is a normal 3.5mm jack having a single break contact which is used to automatically mute the speaker when the earphone is plugged in. Ideally the speaker should have an impedance of about 15 to 26 ohms, but any unit having an impedance in the range 8 to 40 ohms can be used. C4 filters out any RF signal that breaks through to the output and which might cause instability.

The layout of the unit is not too critical, but C2 should be mounted physically close to IC1. As with any set having a ferrite aerial and internal speaker, the two should not be positioned very close to one another as this could cause instability. A metallic case cannot be used as it would shield the aerial and prevent any significant signal pick up. Large metallic components positioned right along side the aerial can have a similar effect.

Short Wave Converter

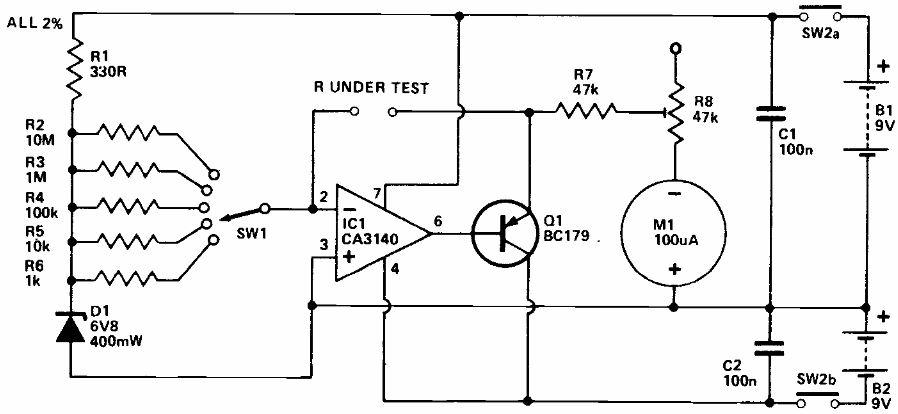


This SW converter tunes over 5 to 15MHz approximately and also enables an ordinary MW broadcast receiver to pick up stations operating on the 19, 25, 31, 41 and 49m broadcast bands.

Signals picked up by the telescopic aerial are directly coupled into the aerial tuned circuit as these signals will be quite weak, necessitating a tight coupling. CV2 is the main tuning capacitor and CV1 is the aerial trimmer control. The signals selected by the tuned circuit are coupled directly into the gate of mixer transistor Q1, no coupling winding being needed here due to the use of a JFET transistor with a very high input impedance. The drain load for Q1 is a MW ferrite aerial, but it is used in reverse in this application, to radiate the 1.6MHz output of the converter. This is picked up by the MW radio, which is placed near the converter and tuned to a quiet spot on the band in the vicinity of 1.6MHz. The position of the coil on the ferrite aerial is adjusted to resonate L2 at the appropriate frequency and effect optimum signal transfer.

The oscillator uses JFET device Q2 in the source follower mode, with positive feedback provided by L3. At the resonant frequency of L3 there is sufficient feedback to cause oscillation and CV3 tunes the oscillator over a frequency range which is 1.6MHz higher than the range of the aerial tuned circuit, so that the required difference frequency of 1.6MHz is produced at the output. C6 is a padder capacitor which gives reasonably good tracking between the aerial and oscillator circuits. Perfect tracking is not required since CV1 can be used to keep the unit peaked for optimum results. C3 is used to couple the output from the oscillator to the input of the mixer stage. The circuit has a current consumption of only 4mA.

Linear Scale Resistance Meter



Although even the most simple of multimeters have resistance ranges, many instruments only have a few ranges, and these have a reverse reading, non-linear scale. This often results in poor accuracy and inconvenience in use. This simple circuit has five measuring ranges from 1kR to 10 Megohms FSD (full scale deflection) with a forward reading linear scale on all ranges.

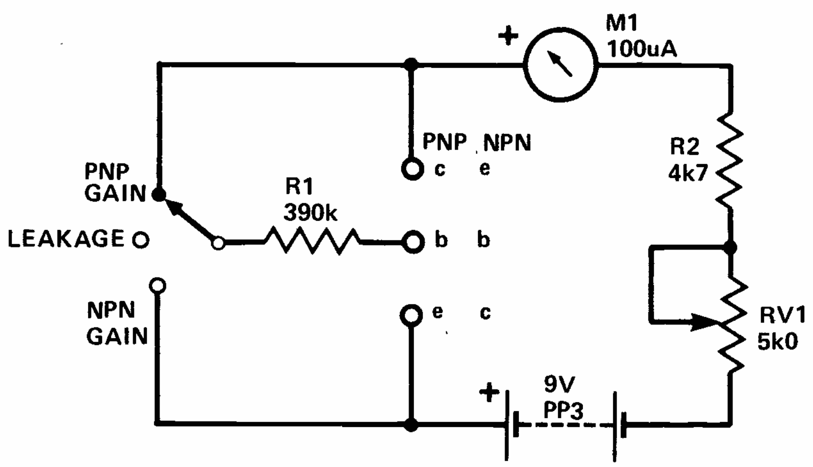
The unit consists basically of an operational amplifier used in the inverting amplifier mode. Transistor Q1 is used as an emitter follower output buffer stage, and on the 1kR range the output sink current capability of the amplifier would be inadequate without the inclusion of this stage. R1 and D1 provide a stable reference voltage of 6V8 (nominal) which is fed to the input of the amplifier. The gain of the amplifier is determined by two resistors, one of which connects the input signal and the inverting input of the op amp. This resistor is one of R2 to R6, depending upon the setting of SW1. The other resistor connects between the amplifier output and the inverting input and, in this case is the resistor under test.

The voltage gain of the circuit is equal to the value of the input resistor divided by the value of the test resistor. Thus, with SW1 switched to the 10kR range for example, a 10kR test resistor would give a voltage gain on one, and the output would swing 6V8 negative. This would give FSD of the simple voltmeter circuit comprised of R7, R8 and M1, which is connected across the output and has a FSD sensitivity equal to the reference voltage. If the test resistor had a value of 5kR, then the circuit would have a voltage gain of only 0.5, and only half FSD of M1 would result. A resistor of 1kR value would give a gain of 0.1 and a deflection of only 10% of FSD. As will be apparent from this, there is a linear relationship between the test resistor value and the meter reading, and the FSD value is equal to that of the resistor selected by SW1.

SW2 is the on/off switch and should be a non-locking push-button switch, or some other type biased to the off position. This is only operated when the resistor has been connected to the test clips as the meter will be deflected beyond FSD if power is applied to the circuit with no test resistor connected (or one of greater value then the FSD value of the range). The meter will not be damaged if this is accidentally done since a maximum meter overload of only about 30% or so can occur.

In order to calibrate the unit, connect a close tolerance resistor, of the same value as that selected by SW1, across the test clips' and adjust R8 for FSD of M1.

Transistor Tester (1)



The cheapest way of building up a stock of transistors is to purchase "bargain bags" from component stores. The main drawback is that you have no idea what characteristics the devices will have, if indeed they have any characteristics at all! Usually between 10% and 30% of the devices will be useless, the remainder will be out of tolerance and not suitable for commercial use – but perfect for use by the experimenter.

With the aid of this transistor tester you will be able to test transistors for gain and leakage characteristics.

To avoid complex switching only one socket is used for the testing of devices – NPN and PNP types are inserted in opposite ways. By comparing the meter deflections caused by good quality transistors, an evaluation of the characteristics of any particular device can be made – high gain, low gain, high leakage, low leakage, etc. Germanium transistors (such as OC71s) will register much higher leakage than more modern silicon types (BC108s, etc.). If no reading is given for either gain or leakage the transistor is open circuit and useless. R2 and RV1 are included to protect the meter movement if the device under test is short circuited internally.

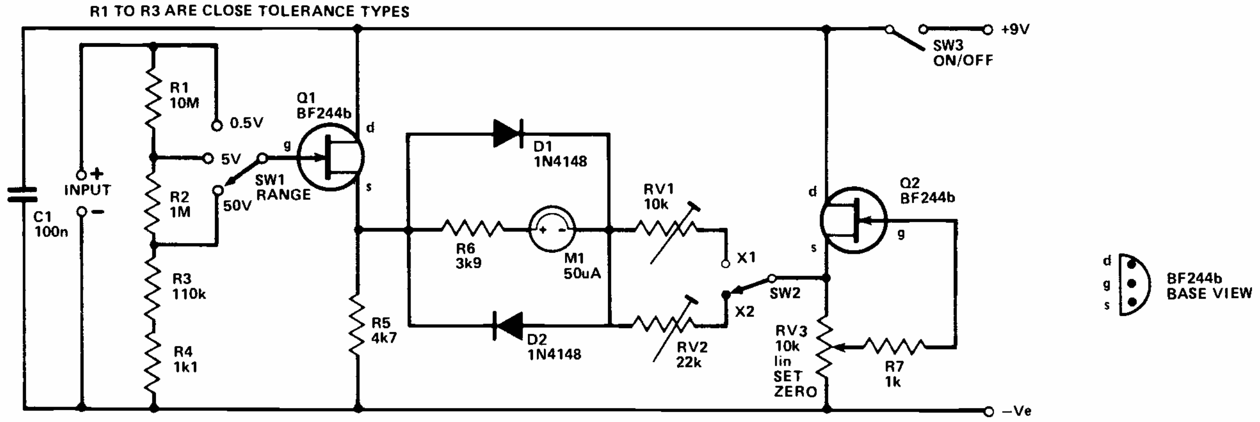
FET Voltmeter

Although an ordinary multimeter is suitable for most DC voltage measurements, it can occasionally prove to be inadequate when making measurements on high impedance circuits which cannot supply the current required to operate a sensitive moving coil meter. The loading effect of the meter then causes the voltage at the test point to fall substantially, giving a misleading reading.

The problem is overcome by this FET voltmeter circuit which has six ranges from 0V5 to 100V FSD, with an input impedance of a little over 11MR on all ranges. This gives a sensitivity of over 22 meg/V on the 0V5 range dropping to a little over 110k/V on the 100V range (most multimeters have a sensitivity of 20k/V).

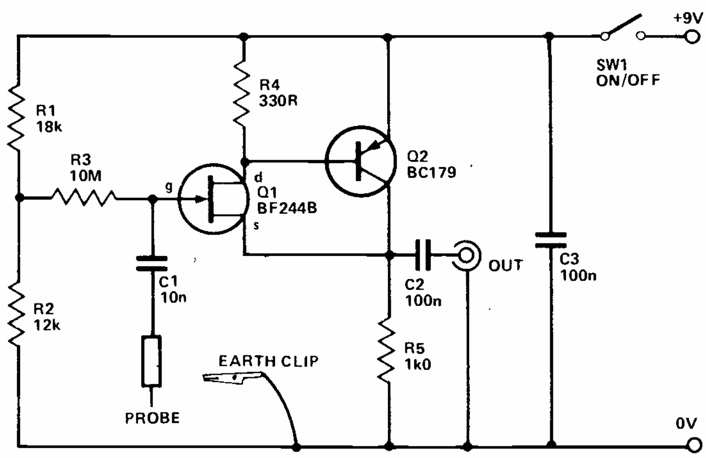
A FET unity voltage gain buffer amplifier based on Q1 is used to give the necessary high input impedance. A simple voltmeter circuit is fed from its output, and this has a FSD value of 0V5 in the 'X1' position of SW2, or 1V in its 'X2' position. RV1 and RV2 respectively are adjusted to give the circuit the correct FSD values. There is a small quiescent output voltage from the buffer stage and so a bridge circuit is used to give zero quiescent voltage across the meter circuitry. To give good stability another source follower is used to form the other section of the bridge and this results in no noticeable meter drift. RV3 is used to electrically zero the meter. D1 and D2 simply protect against serious over-loading.

An input attenuator can be used to reduce the basic sensitivity by a factor of 10 or 100, giving FSD values of 0V5, 5V, and 50V with SW2 in the 'X1' position, or 1V, 10V and 100V if it is set to the 'X2' position.



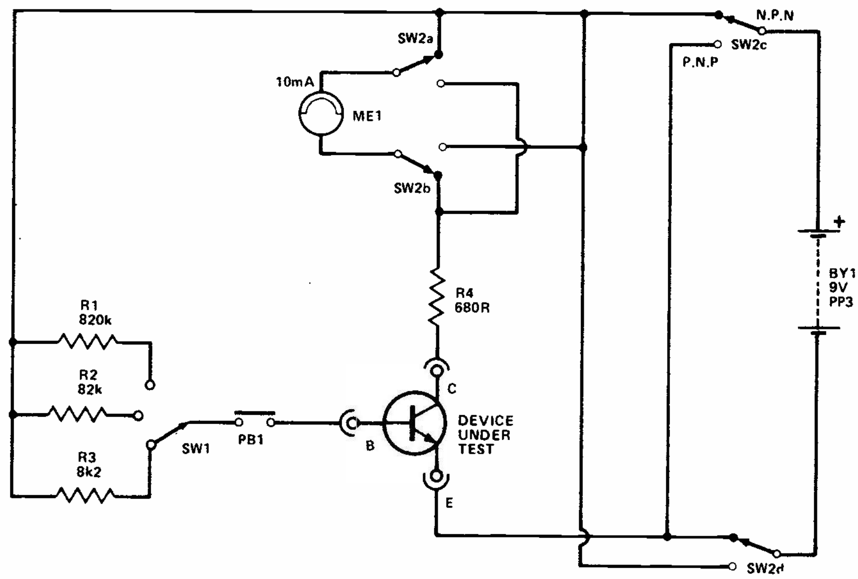
Active 'Scope Probe

Most oscilloscope probes that provide a tenfold boost in input impedance (usually from 1MR to 10MR) are passive devices and consequently give a corresponding tenfold reduction in voltage gain. This active probe circuit gives an input impedance of approximately 10MR shunted by about 5 to 10pF and has a voltage gain of almost exactly unity. It will not have any significant effect on the bandwidth when used with an oscilloscope whose bandwidth is 10MHz or less and which is AC coupled (the probe is mainly intended for use when investigating high impedance audio or RG signals). Signal levels of up to about 3Vrms can be handled without clipping or about 6Vrms if the supply voltage is raised to 18V.



The obvious choice for a circuit of this type would at first seem to be a FET used in the source follower mode as this can provide a high input impedance, low input capacitance, low output impedance and nominally unity voltage gain. The problem with such a circuit is that the true voltage gain is only about 0.9 to 0.95, which would obviously have an adverse effect on the accuracy of any form of calibration fitted to the 'scope. Therefore, the circuit finally devised uses Q1 as a common source amplifier, which directly drives common emitter amplifier Q2. There is a 100% negative feedback loop from the collector of Q2 to the source of Q1 and this gives the circuit almost precisely unity voltage gain. This arrangement retains the attributes of a source follower stage and gives good results in other respects as well. Gate biasing for Q1 is provided by R1-3 while DC blocking at the input and output is provided by C1 and C2 respectively. Current consumption of the unit is about 5mA.

Transistor Tester (2)



This very simple transistor tester measures DC current gain in three ranges, with full scale values of 10, 100 and 1,000. It will also show whether or not the device under test has a high leakage current.

The basic tests for a transistor are very simple and in order to test for high leakage, it is merely necessary to connect a voltage across the emitter and collector terminals of the test device, and then measure the current flow.

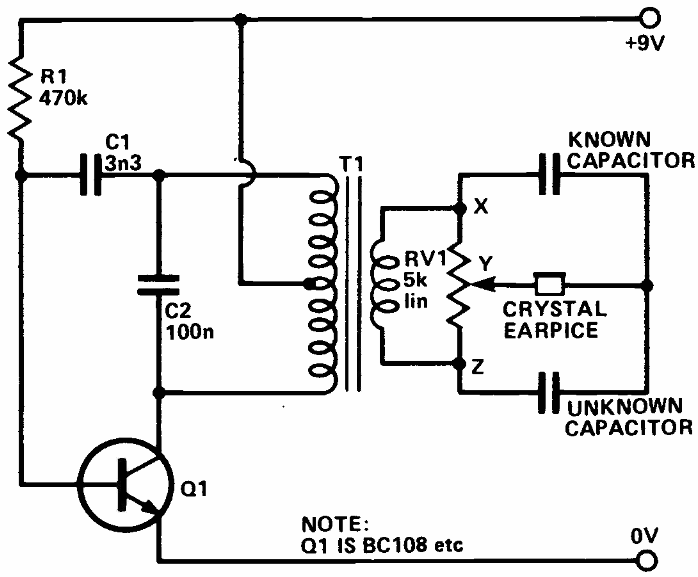
In this circuit BY1 is the voltage source, and ME1 registers the current flow. R4 is a current limiting resistor which protects ME1 and the test device from passing an excessive current.

Silicon devices have extremely low leakage currents, and if there is any deflection of ME1 when testing a silicon transistor it certainly means that device is not functional (or is connected incorrectly). Germanium devices have somewhat higher leakage currents, and a very small deflection of ME1 is acceptable.

The test for DC current gain (Hfe) is basically the same as for leakage testing, except that a current is fed into the base terminal of the test device. This causes a larger current to flow in the collector circuit of the transistor, and the current gain is equal to the collector current divided by the base current. If SW2 is depressed, a base current will be provided to the test device by which ever resistor (R1 to R3) is selected by SW1. With SW1 in the '10' position, a nominal base current of 1mA is fed to the test device, and it must have a current gain of 10 in order to produce a collector current of 10mA and give full scale deflection of ME1. Lower levels of current gain give a proportionately lower meter reading. With SW1 in the '100' and '1,000' positions, the base current is reduced to 100μA and 10μA respectively, giving the correspondingly higher full scale gain values.

PNP and NPN transistors require opposite supply polarities, and SW3 is used to switch the supply polarity to suit the type being tested and to connect ME1 with the correct polarity.

Measuring Capacitance



This simple circuit will enable you to find out the value of an unknown capacitor. It makes use of the Bridge principle.

The interesting part of the circuit is to the right of T1; the rest, including Q1 is simply an audio oscillator. T1 is a small output transformer, as found in nearly all transistor radios. A commercial version is widely available from component shops (LT700). Connected as shown, the circuit forms a Hartley oscillator. C2 converts the primary of the transporter into a tuned circuit operating in the audio range whilst C1 feeds back part of the signal.

Let us take a case where the two capacitors are the same value and the resistance in RV1 between x-y and y-z is the same; in that case the voltage at y and at the junction of the two capacitors will be the same and nothing will be heard in the earpiece.

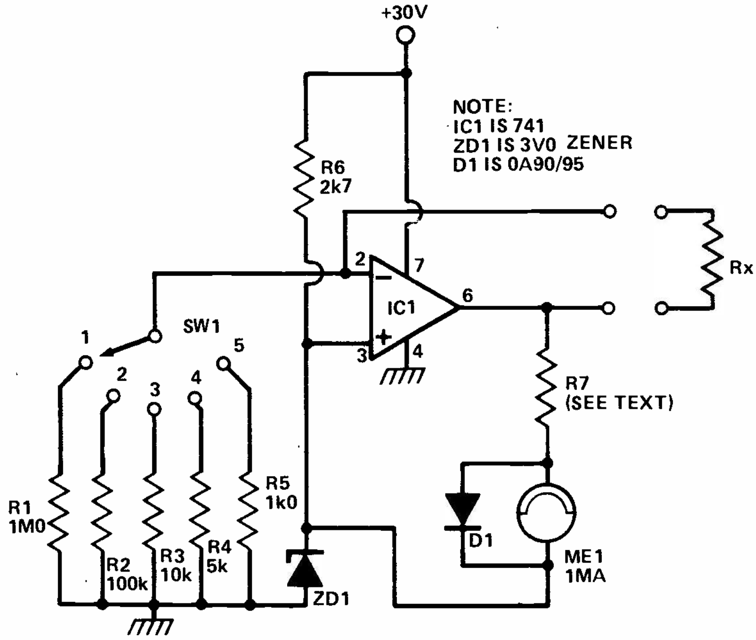
Assume now that our unknown capacitor is half that of our known component. A larger amount of the signal will pass through the known capacitor, and a signal will be heard in our earpiece. However if x-y is twice y-z, balance will once again be achieved and nothing will be heard. It follows that if the knob of RV1 is marked in ratios, we will be able to calculate the value of almost any capacitor as long as we use a reference component that is between ten times and one-tenth of the unknown this is because it is only practical to mark out ratios of 10:1. This is not as much of a problem as may first be imagined, as values between 1p and 100uF can be checked using four standards – these are 10p, 1n (1000pF), 100n (0u1f) and 10uF.

Direct Reading Ohmmeter

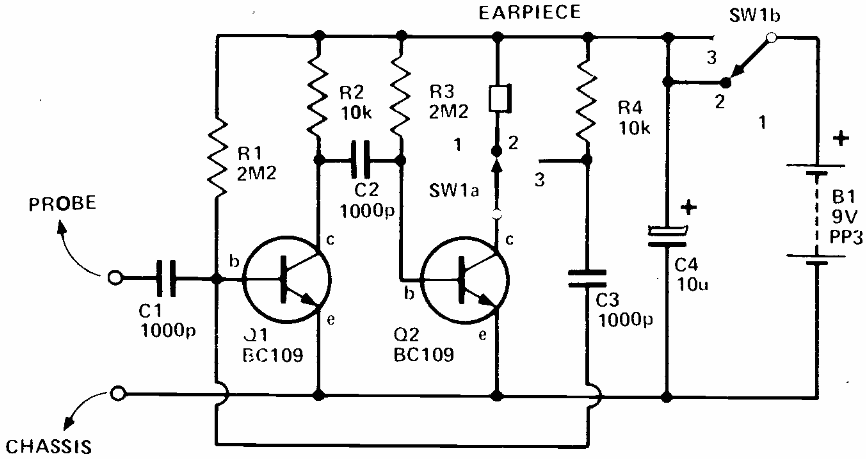
This is a useful piece of epuipment for checking unknown resistors, continuity of coils, transformer windings etc. Designed around a 741 operational amplifier it has the advantage that once the meter dial is calibrated, resistance can be read directly from the scale without the usual cramping at the low-resistance end.

The meter can have any internal resistance, but it must be made up to 3k by an external resistor R7. The prototype used a meter with an internal resistance of 100R. It is a good idea to make up the meter resistance to a round value of say, 300R, by a small resistance so that you aren't left with an awkward value for R7.

The complete unit is self-zeroing and is protected against open-circuit or unknown resistance at Rx by diode D1.



Signal Injector-Tracer



There are two extremely useful pieces of test gear for both the serviceman and the amateur constructor: these are a signal source and a signal tracer.

Faced with a transistor radio that doesn't work, what do you do? It is important that a logical approach is taken and although this may sound obvious, it is very, very easy to become diverted.

First check that the battery is not flat (this accounts for about 50% of so called faults) and then check that a good contact is being made on the cut-out switch of the earpiece socket if one is fitted. Always chech these first but assuming there is still no joy, what do you do?

The volume control is easily located, contacts can generally be made to it quickly and it is an excellent place to start.

If you inject a signal into the slider of the volume control and it is heard at a decent level you can be fairly sure that nothing is wrong with the amplifier. If nothing is heard there is obviously something wrong and the field is immediately narrowed.

Assuming that the audio stage is working you can then inject an IF signal at the collector of the mixer stage – the same rules apply as before.

Alternatively you can take the 'signal detect' approach. Instead of injecting a signal at the volume control you can listen at the same point to establish that the radio is working satisfactorily.

The above is a super-concise lesson in fault finding but it does illustrate the tremendous use that a signal injector and a signal tracer can be put to.

The project described here is for a combined device – it can inject signals at RF, IF and AF and can detect signals at the same frequencies, assuming that they are high enough in level. The simplicity of circuit may lead you to doubt this claim but it does do all this. The function switch, SW1, has

No. 1	Off position
No. 2	Tace Position
No. 3	Inject Position

Position 1 merely disconnects the supply and the device is of course inoperative. As shown the function switch is in position 2, the trace mode.

One of the contacts is the common line and should be wired using a crocodile clip to the chassis of the equipment being investigated. The other connection is the probe.

This goes via DC blocking capacitor C1 whose working voltage should be high; if a 500V component is used the circuit can be used on valve equipment.

The signal is fed to Q1 which is arranged as a common emitter amplifier but is biased nearly to cut-off, which creates deliberate distortion at the same time as amplifying the signal. Distortion in such a manner leads to the detection of RF signals and so whatever the frequency fed in, assuming it is modulated an audio output will be heard. The collector load of Q1 is R2 and the output of this stage is fed to another of similar design, but the load is a high impedance magnetic ear-piece.

On inject, SW1 is in position 3 and the output of Q2 is coupled to R4, acting as the collector load, and also to C3 which feeds back to the base of Q1. The circuit, which was previously an amplifier, now becomes a multivibrator producing a square wave signal at approximately 1kHz and this is fed, via C1, to the probe.

A square wave can be described as a fundamental frequency plus all its harmonics and so in addition to 1kHz there is an output at 2kHz, 3kHz etc., going right up into the RF range. In fact, there is still a useable output at 30MHz.

Holding the probe near the aerial will produce an output from a working radio, as the injector is working as a very low power transmitter and a 1kHz tone will be heard from the loudspeaker.

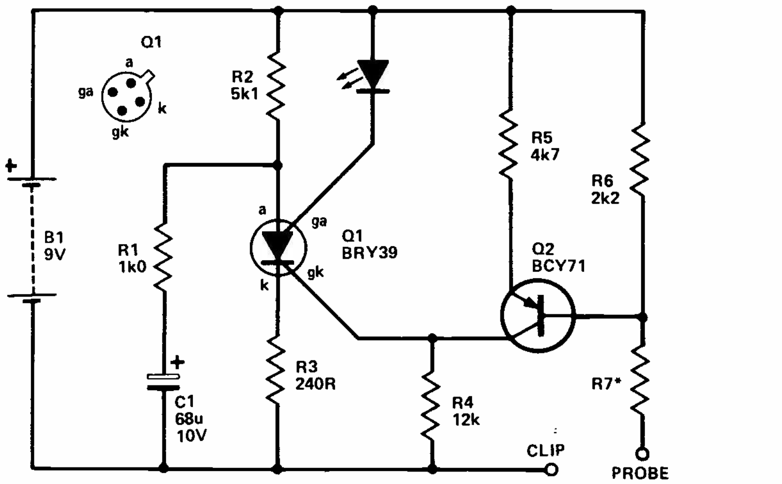
High gain transistors are needed in order to hear really low signal sources and high frequencey types are needed to handle the upper harmonics. A transistor incorporating both these qualities is the BC109.

The current drain, both in the trace and inject mode is quite small. SW1, the function switch, needs to be a 2-pole, 3-way rotary switch.

Note that only high impedance magnetic earpieces are suitable, though 2000 ohms headphones can be used instead.

Once completed the signal injector/tracer will be found to be almost indispensable and for this reason it is worthwhile building the circuit carefully and neatly into a small chassis.

Suppressor Tester



Few motorists think of checking ignition leads or caps, yet these resistive components can be notoriously unreliable.

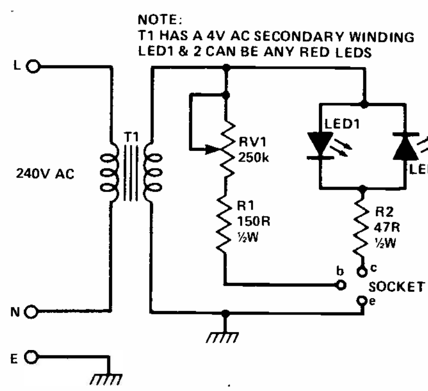
For adequate interference suppression (required by law) the resistance between distributor cap and spark plug must be at least 10k. However, for most efficient operation the resistance should not exceed 20k. Also, any open circuit will allow arcing, causing high levels of interference and rapid deterioration of the suppression components.

The circuit is based upon a programmable unijunction transistor (PUJT) oscillator which flashes a red LED when the suppressor resistance is correct. The oscillator will function only when the voltage on the cathode-gate of Q1 is within its operating range. This voltage is determined by the current flowing through Q2, which is controlled by the suppressor resistance connected across the input wires of the tester. Since the position of the working range is determined by component tolerances and by the gain of Q2 it is necessary to select a value for R7 (between 100R and 10k).

With an 8k2 resistor across the input wires, D1 should give a steady light. With 22k it should not light. Alter the value of R7 until the range of suppressor values which allows D1 to flash is centred about 15k. The probe can be made from a brass screw and an old spark plug top. A crocodile clip is used to make the connection inside the distributor cap. No on/off switch is required as current drain is negligible with D1 'off'.

WARNING: the engine *must not* be running while the supressor components are being tested.

Diode-Transistor Tester



There is nothing particularly new about this circuit but it is very useful when checking diodes and transistors and can save a great deal of time.

Potentiometer RV1 is kept at maximum resistance when a transistor is plugged into the socket, then slowly decreased until one of the LEDS glows brightly, depending on the type of transistor (NPN or PNP). If both LEDs light up, the transistor has a short-circuit: if neither LED lights the device has an open-circuit.

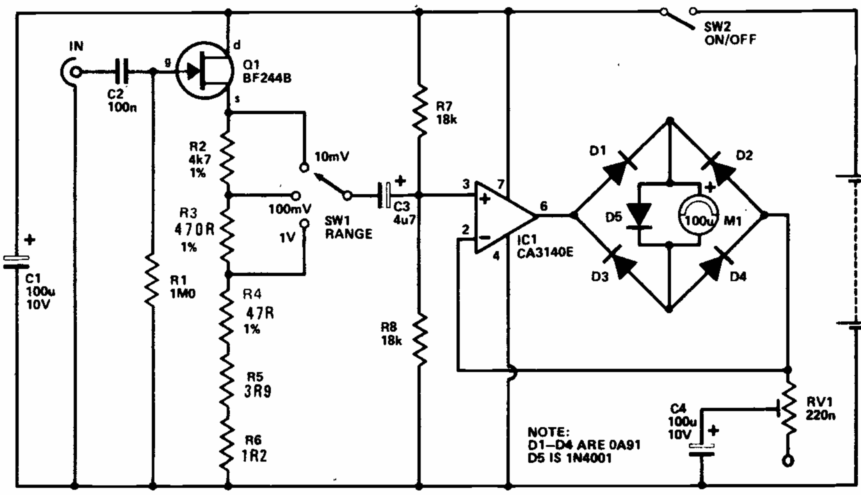
The same method applies to diodes. If the two leads of the diode are plugged into E and C (anode to E) the PNP LED will light up. If reversed the other LED will glow. As with transistors, if both LEDs light the diode is open-circuit.

By plugging in a new transistor with a known gain an observing the brightness of the LED and the position of RV1, you will get a rough idea of the gain of unknown devices.

AF Millivolt Meter

This simple and inexpensive millivolt meter has three measuring ranges of 10mV, 100mV, and 1 Vrms full scale sensitivity. The frequency response has −1dB. points at about 20Hz and 75kHz. The instrument is suitable for making audio noise, frequency response, and gain measurements, and would be useful to any beginner interested in audio.

The unit uses a conventional arrangement with a non-inverting op-amp circuit feeding a meter via a bridge rectifier. The negative feedback loop is taken to the inverting input by way of the rectifier and meter circuit rather than direct from IC1 output. With low voltages applied to the rectifiers they have a high forward resistance, but this results in little feedback and the amplifier having a high gain. Thus small input signal amplitudes which would otherwise produce no meter deflection due to the high rectifier resistance are boosted to the point where they give the appropriate meter reading. Therefore, although the rectifier is inherently non-linear, it produces opposing non-linear feedback which compensates for this and gives the unit linear scaling. RV1 is used to adjust the circuit to the correct sensitivity and D5 protects the meter against severe overloads.



Q1 is used as a low noise, source follower buffer amplifier which gives the circuit a high input impedance, about 1Meg. This ensures that the instrument imposes little loading on the equipment under test. An attenuator is incorporated at the output of the buffer stage, and this can be used to reduce the basic 10mV sensitivity to 100mV or 1V FSD. The attenuator does not need any frequency compensation as it is in a low impedance part of the circuit.

To calibrate the unit it is switched to the 1V range, RV1 is set at maximum and with a 1V rms audio source connected to the input, RV1 is adjusted for full scale deflection of the meter. The 1V audio source can be provided by an AF signal generator set to the correct output level with the aid of a multimeter switched to a low AC volts range.

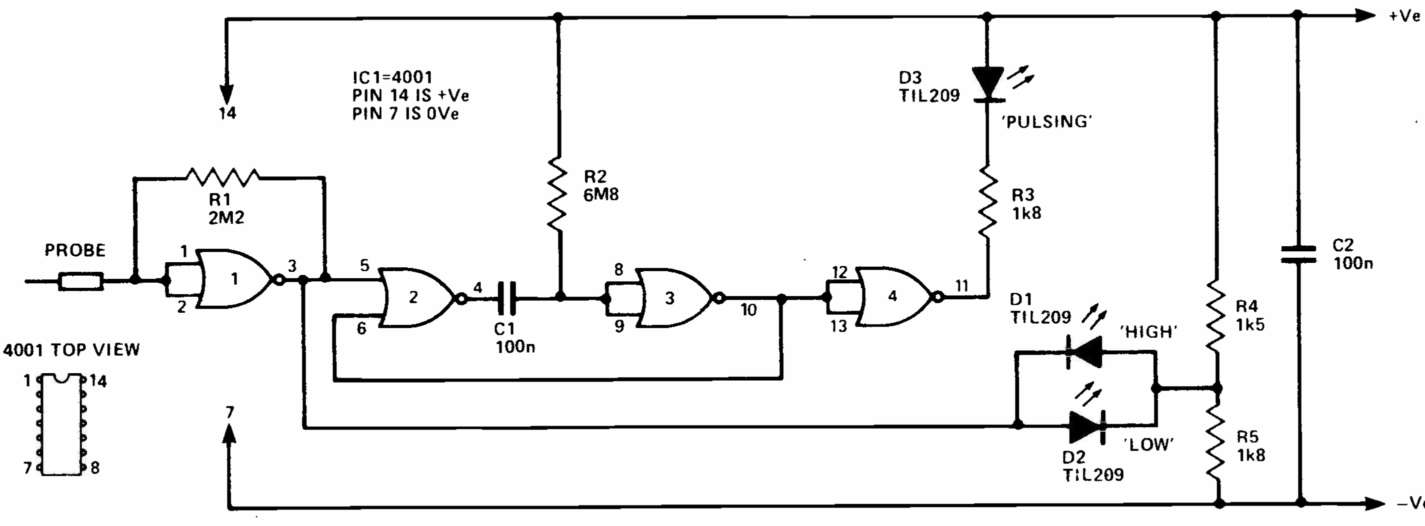
CMOS Logic Probe

A logic probe is a device used when testing digital circuits, and it shows the logic state at the selected test point. This one can indicate four states, as follows:

1. Input high (logic 1).
2. Input low (logic 0).
3. Input pulsing.
4. Input floating.

The circuit uses the four 2 input NOR gates contained within the 4001 CMOS device, and is primarily intended for testing CMOS circuits. The probe derives its power from the supply of the circuit being tested. The first gate has its inputs tied together so that it operates as an inverter, and it is biased by R1 so that roughly half the supply potential appears at its output. A similar voltage appears at the junction of R4 and R5, and so no significant voltage will be developed across D1 and D2. Thus under quiescent conditions, or if the probe is connected to a floating test point, neither D1 nor D2 will light up. If the input is taken to a high logic point, gate 1 output will go low and switch on D1. If the output is taken to a low test point, gate 1 output will go high and D2 will be switched on.

A pulsed input will contain both logic states, causing both D1 and



D2 to switch on alternately. However, if the mark space ratio of the input signal is very high this may result in one indicator lighting up very brightly while the other does not visibly glow at all. In order to give a more reliable indication of a pulsed input, gates 2 to 4 are connected as a buffered output monostable multivibrator. The purpose of this circuit is to produce an output pulse of predetermined length (about half a second) whenever it receives a positive going input pulse; the length of the input pulse has no significant effect. D3 is connected at the output of the monostable, and is switched on for about half a second whenever the monostable is triggered, regardless of how brief the triggering input pulse happens to be. Therefore a pulsing input will be clearly indicated.

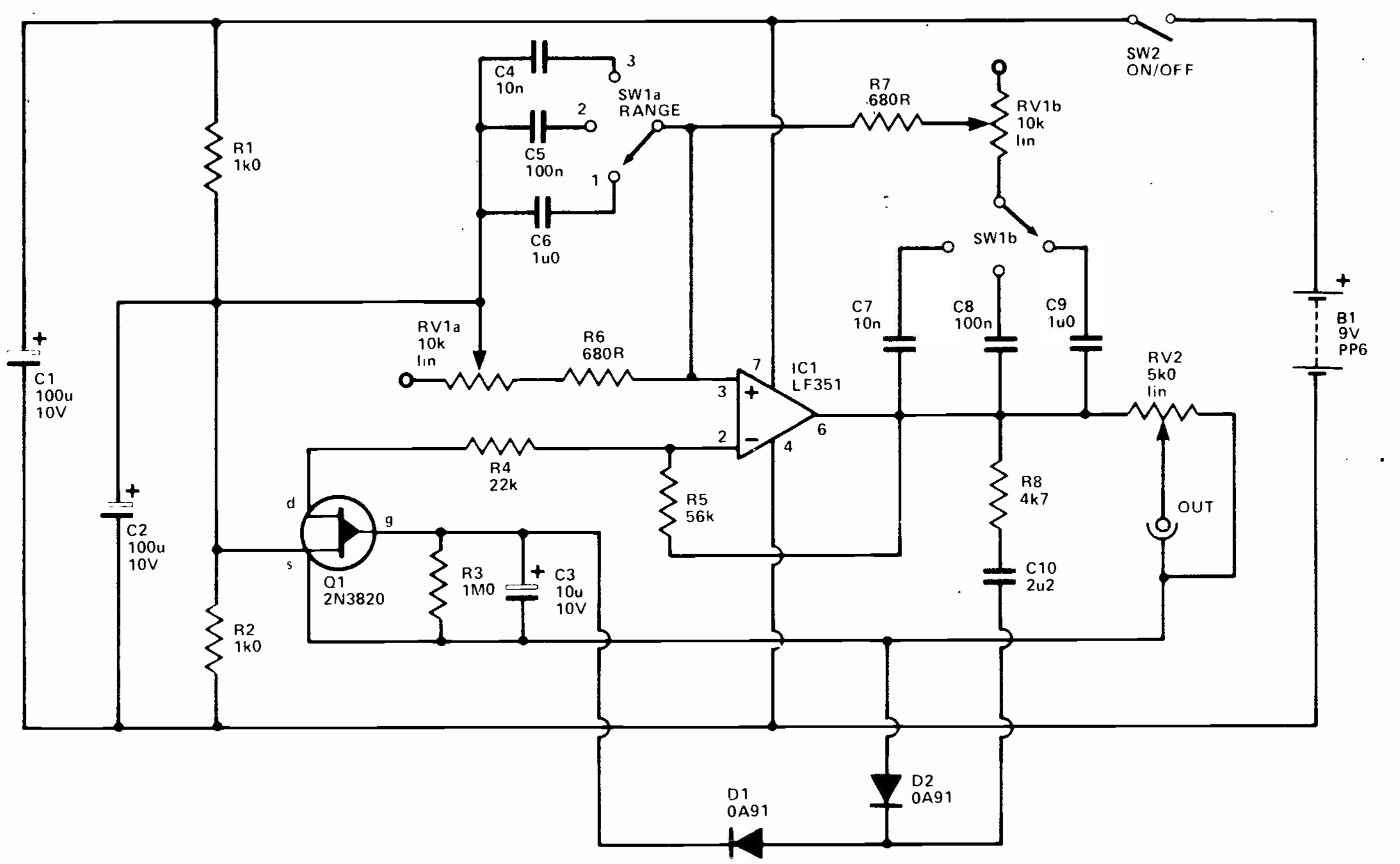
A. F. Signal Generator

The circuit shown here provides a good quality sinewave output over three continuously variable ranges (Range 1, below 20Hz to above 200Hz; Range 2, below 200Hz to over 2kHz; and Range 3, below 2kHz to over 20kHz) covering more than the entire audio frequency spectrum.

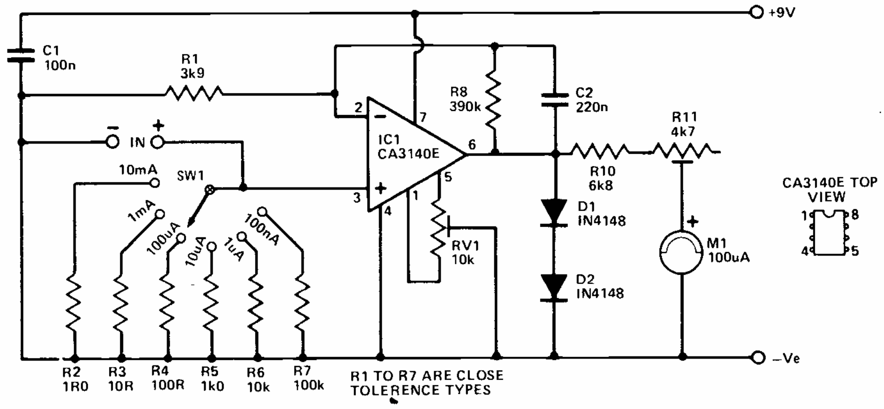
The circuit uses the usual Wien Bridge type circuit, which consists of an amplifier having frequency selective positive feedback provided via a C-R network. The capacitive elements of this network are whichever two capacitors are selected by SW1. The resistive elements are R6, R7 and RV1, the latter permitting the unit to be tuned over the ranges quoted above. This network provides positive feedback over operational amplifier IC1, which is a FET type giving low noise and distortion levels. RV1a and R6 also bias the non-inverting input of IC1 to a central tapping on the supply produced by R1, R2 and C2.

The closed loop gain of IC1 must be maintained at precisely the correct level if good results are to be attained. An automatic gain control (AGC) circuit is used to maintain stable operating conditions and a constant output level. R5, R4 and the drain to source resistance of Q1 form a negative feedback network which controls the closed loop gain of IC1. Initially Q1 is forward biased by R3 so that there is enough gain to give strong oscillation. Some of the output from IC1 is coupled by R8 and C10 to a rectifier and smoothing network composed of D1, D2 and C3. These produce a positive bias which tends to cut off Q1, producing reduced circuit gain. The stronger the circuit oscillates, the larger the bias, and the lower the gain. Lack of oscillation produces reduced bias, more gain, and stronger oscillation.

Variable attenuator RV2 enables the output to be adjusted from zero up to about 1V5 rms. The current consumption of the circuit is about 7 mA.



Nanoamp Meter



It is not possible to accurately measure currents of a few microamps or less using an ordinary panel meter or multimeter. In order to make such measurements it is necessary to use an active circuit such as the one shown here. It can be built as a self-contained unit or used as part of an instrument requiring a highly sensitive current meter. The sensitivity is from 100nA to 10mA in six ranges, the higher ranges being included to permit calibration, and because many multimeters have very few low current ranges.

M1 is connected in a 1V FSD voltmeter circuit which also uses R10 and R11. The latter is adjusted to give the unit the correct sensitivity. IC1 is an op amp connected in the non-inverting mode and having a DC voltage gain of about 100 times (set by feedback network R8-R1). C2 reduces the AC gain to about unity so as to improve stability and imunity to stray pick-up. The non-inverting input of IC1 is biased to the OV rail by whichever of the range resistors (R2-R7) is selected by SW1. In theory this gives zero output voltage and no meter deflection, but in practice it is necessary to compensate for small offset voltages using offset null control, RV1.

If an input current is connected to the unit, a voltage will be developed across the selected range resistor, this voltage being amplified to produce a positive meter deflection. With R2 switched into circuit. 10mA is needed to give full scale deflection since 10mA will cause 10mV to be developed across R2. This will be amplified one hundred fold by IC1 to give one volt at the output. On successive ranges, the range resistor is raised by a factor of ten, reducing the current required at the input to develop 10mV and give full scale deflection of M1.

This arangement relies on the amplifier having a very high input impedance so that it does not drop a significant amount of input current, and this is achieved by using a FET input op amp having a typical input resistance of 1.5 million meg ohms. D1 and D2 prevent the output voltage of IC1 from exceeding more than about 1.3 volts, and they thus protect M1 against overloads.

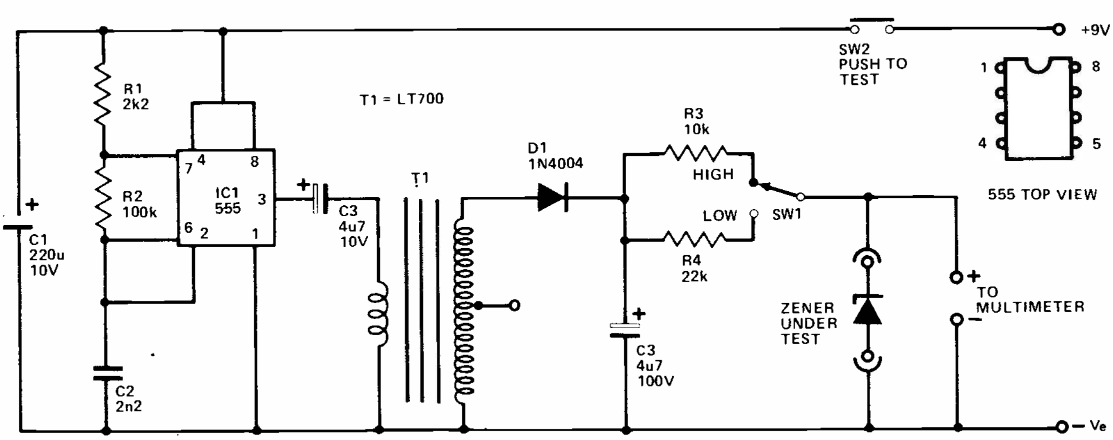
When adjusting RV1 start with its slider at the pin 5 end of the track (there should be a strong deflection of M1), and then back it off just far enough to zero the meter, and no further.

Zener Diode Tester

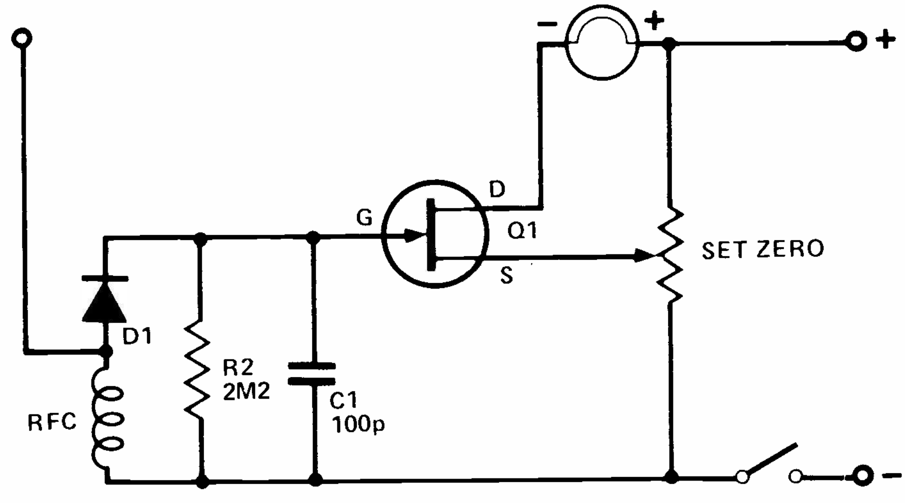
This circuit is an add-on unit for a multimeter having a sensitivity of 20k/V or better, and it enables a rough check to be made on zener diodes having operating voltages of up to about 33 volts. The unit operates from a standard 9 volt battery (PP6, PP7, or PP9 size).

In order to obtain a suitably high voltage for this application from an ordinary 9 volt dc supply it is necessary to have a voltage step-up circuit of some kind. An audio oscillator using IC1 is used to drive the primary winding of step-up transformer T1, giving about 50Vac from the secondary winding. The output from T1 is halfwave rectified and smoothed by D1 and C3 to give to give an unloaded DC supply of about 75 to 80 volts (about 40 to 50V when loaded).

With SW1 at the 'low' position, a current of about 1 to 2mA. (depending upon the voltage of the zener under test) is fed to the test device through current limiting resistor R4. When SW2 is operated the multimeter, which is switched to an appropriate DC voltage range, is connected in parallel with the test device and registers its zener voltage. Switching SW1 to the 'high' position doubles the current to the zener under test, as a lower value current limiting resistor (R3) is then switched into circuit. If the test divice is fully functional this should cause only a very small increase in the meter reading, and there may well be no noticeable change.



Field Strength Meter



The device will operate at any frequency up to 250MHz or even higher if necessary. A short whip, rod, telescopic or other aerial picks up radio frequency energy and rectification by diode D1 provides a positive voltage for the FET gate, across R1. This FET is only operating as a DC amplifier and the 2N3819 or other general purpose transistor will be satisfactory.

The "Set Zero" potentiometer may be 1k to 10k. With no RF signal present, it allows gate/source potential to be adjusted, so that the meter shows only a small current, which rises in accordance with the strength of the RF present. For high sensitivity, a 100uA meter can be fitted. Alternatively, a meter of lower sensitivity, such as 250uA, 500uA or 1mA can be used and will provide enough indication in most circumstances.

Should the field strength meter be wanted for VHF only, a VHF choke can be used, but for general usage over lower frequencies, a short wave choke is necessary. An inductance of about 2.5mH is satisfactory for 1.8MHz and higher frequencies.

The device can be constructed in a small insulated or metal box, with the aerial projecting vertically. In use, it allows tuning up a transmitter final amplifier and aerial circuits, or the adjustment of bias, drive and other factors, to secure maximum radiated output. The effect of adjustments will be shown by the rise or fall of the reading of the meter.

AC Meter Booster

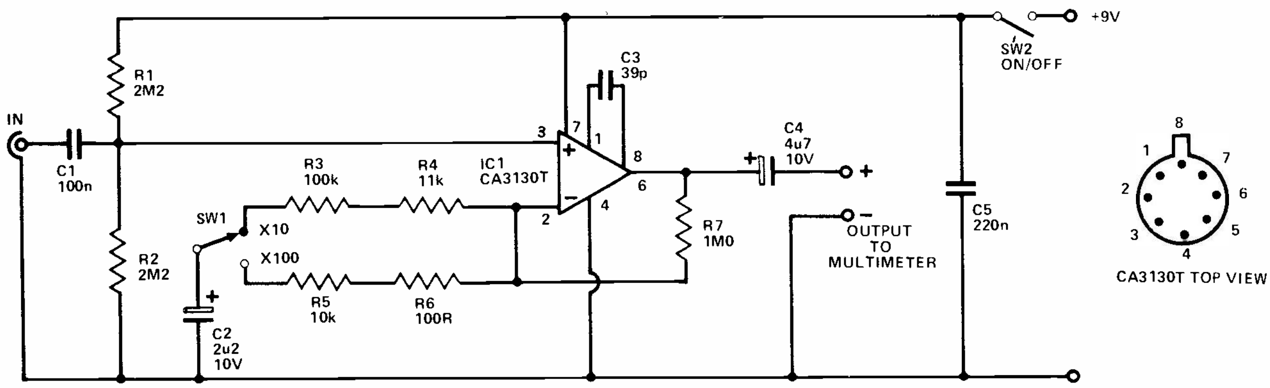
Measuring small audio frequency signals is often impossible, using an ordinary multimeter, because most of these have a low AC range of about 1 to 5V FSD. A simple and inexpensive solution to the problem is to add an amplifier ahead of the multimeter. This amplifier has a switched voltage gain of 10 or 100, and would therefore boost the sensitivity of (say) a multimeter switched to the 2V5 AC range to 250mV and 25mV FSD respectively. measurements down to just a few mV rms can then be made with reasonable accuracy.

The circuit uses a CA313OT operational amplifier in the non-inverting mode. The non-inverting input is biased to about half the supply voltage by R1 and R2, and the input signal is coupled to this point by C1. The input impedance of the circuit is set at over 1M by R1 and R2, so that the unit places little loading on the circuit under test. R7 biases the inverting input and gives a quiescent output voltage of about half the supply potential. The voltage gain of the amplifier is set by the ratio of two resistances; with SW1 in the 'X 10' position the two resistances are R7 and R3 + R4. The voltage gain is equal to the sum of the resistances divided by the shunt resistance (R3 + R4); that is,

$$G = \frac{R7+R3+R4}{R3zR4}$$

This gives almost exactly the required figure of 10, with the specified values. With SW1 in the 'X 100' position the lower resistance of R5 and R6 is switched into circuit, boosting the voltage gain to almost exactly 100.

DC blocking at the output is provided by C4. C5 is a supply decoupling capacitor and should be mounted physically close to IC1. C3 is the compensation capacitor for IC1, and prevents the device from becoming unstable. Note that a carefully designed layout having the input and output well isolated from one another is required, or the circuit as a whole may become unstable; screened input and output cables should be used. It should be used with the multimeter set to a range of 3 V or less. The amplifier has a flat response up to about 30kHz in the 'X 100' mode, and up to about 300kHz in the 'X 10' mode.



R3, R4, R5 AND R7 ARE 2% OR BETTER

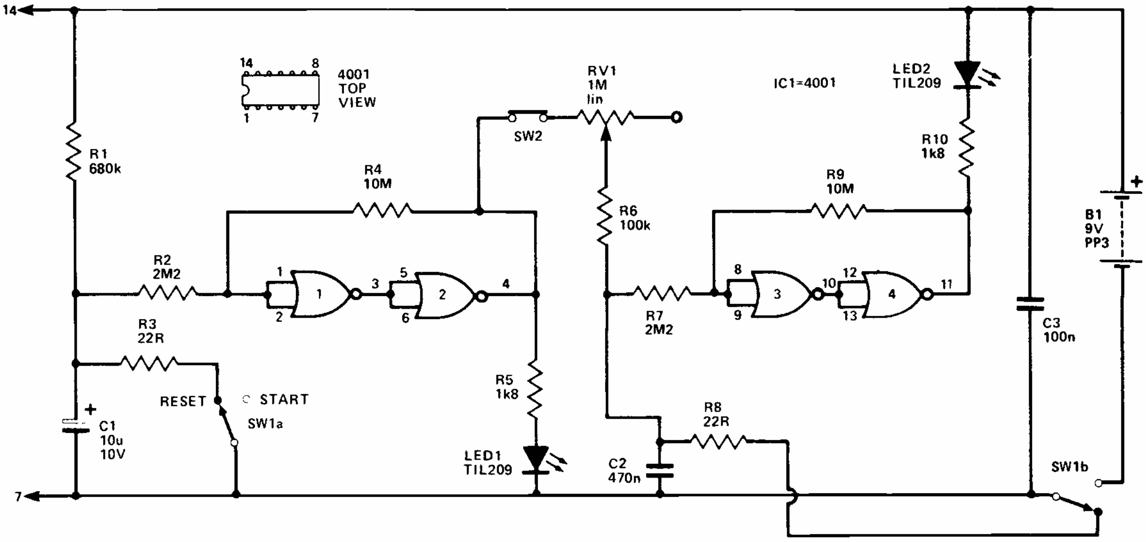
Reaction Game

This is a simple reaction testing game for one player. The idea of the game is to end up with both LED indicators switched on; initially only one lamp will come on, followed by the second one shortly afterwards. When this happens a switch must be operated as quickly as possible, and the first LED will switch off if the attempt is too slow.

The circuit uses the four 2 input NOR gates of 4001 CMOS device, the gates being connected as simple inverters. Gates 1 and 2 are connected to form a Schmitt trigger circuit and at switch-on C1 will be discharged causing gate 1 input and gate 2 output to be low. After about seven seconds C1 will have charged through R1 to the transition voltage of gate 1, with the result that gate 2 output swings positive. Coupling through R4 causes gate 1 input to be taken further positive, and the regenerative action results in gate 2 output jumping to the high state and switching on D1. This indicates that the player should operate the push button switch SW2.

D2 is normally on but will be switched off at the end of the second timing period, unless SW2 is used to halt the charging of C2 in time. The second delay can be adjusted, using RV1, from more than 500ms at maximum resistance to about 50ms at minimum resiatance. This gives a difficulty factor varying from "easy" to "impossible" for anyone with normal reactions.

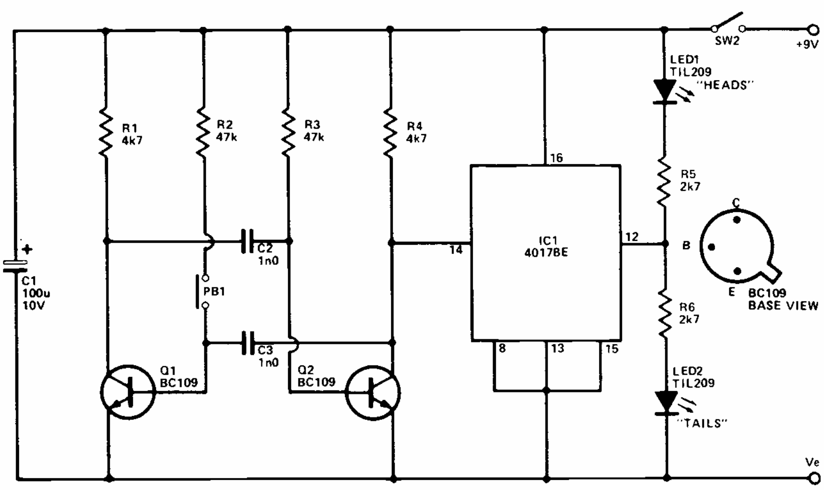
The circuit is reset using SW1 which disconnects power from the circuit, discharges C2 through current limit resistor R8 and similarly discharges C1 through R3. The unit is then ready for operation again when SW1 is set back to the "start" position.



Heads or Tails

This simple novelty circuit is designed to electronically simulate the tossing of a coin. The unit has a push button switch which is briefly depressed in order to "toss the coin"; only one of the LEDs will be switched on when this switch is released.

The circuit uses Q1 and Q2 in a standard astable multivobrator circuit. As the circuit stands there is no bias to Q1 and the circuit therefore fails to oscillate. However, if SW1 is operated the circuit can function normally. A roughly squarewave output is then produced at the collector of Q2, with the specified values give an operating frequency of many kilohertz.



This squarewave output is fed to a 4017 divide by ten circuit; after each five input cycles, the output (pin 12) changes state, and while the clock oscillator is functioning this output changes state a few thousand times per second. D1 and D2, the two LED indicators, are driven from the output via current limiting resistors R5 and R6. When IC1's output is low, R6 and D2 are effectively short circuited by the output stage, but D1 will be switched on. Conversely, when the output is high, D1 and R5 are short circuited, and it is D2 that is switched on. While the oscillator is running both LEDs appear to be switched on. There is of course, no way of predicting which of the LEDs will be switched on. It is purely a matter of chance.

SW2 is the on/off switch. The current consumption of the circuit is 5mA.

Magic Candle

Electronic party tricks are always popular. The majority of people have very little understanding of electronics and even simple tricks can mystify them. This circuit uses only a handful of common components and can be built very quickly. However, as with many projects of this type, the ingenuity in building is probably more important than the circuitry; this is left to the reader, though some general tips are given later.

The idea of the "Magic Candle" is to demonstrate that lightbulbs can be lit by a match or cigarette lighter and can be snuffed in a similar way. The bulb should be the only item that is actually showing but it is important that the LDR – light dependent resistor – is very close by with the active face pointing at the bulb. When a match is struck and brought up to the bulb this causes light to fall on the LDR. The resistance of the LDR falls considerably and since this forms a potential divider with RV1, which is coupled to the base of the transistor, the voltage on the base rises and causes Q1 to conduct.

When the match is withdrawn the light from the bulb takes over as the source that keeps the resistance of the LDR low and so the transistor will remain on and the bulb will stay alight. If now the bulb is "snuffed", by breaking the path of light between the bulb and the LDR, the bulb will go out and remain so until the light level once again reaches sufficient brightness to turn the transistor on.

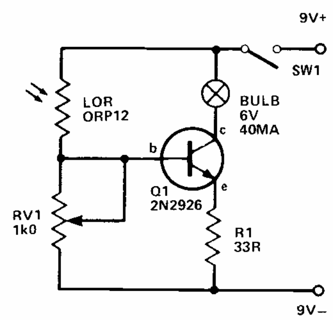
The use of a 6V bulb is simply because these types are widely available and cheap. In order to prevent too high a voltage being applied, resistor R1 is connected in the emitter circuit. In the conducting state there is only a tiny voltage drop across Q1 but about 3V will be dropped across the resistor, thus ensuring that the bulb is not overdriven.

Since the circuit will have to operate in widely differing light levels, it is necessary to control the sensitivity of the circuit and this is accomplished by RV1. In high ambient light levels the value of RV1 should be low, this means that the transistor will remain switched off until the light level created by the match goes above this level. In low light levels the value of RV1 will be high.

RV1 can take the form of a miniature preset control which for normal uses can probably be left at some level found experimentally for general purpose use. It is not possible to give even an indication of this value as the resistance of light dependent resistors varies considerably with individual specimens.

The current drain is 40mA which is rather heavy for a PP3 battery, though one in good condition will work for a short period. The heavy current drain may be acceptable as the circuit is unlikely to be on for long periods and this battery has the advantage of being small in size and cheap. SW1, the on-off switch, can take any convenient form or it may be omitted, the circuit being switched off by removing the battery clips.

As we mentioned before, the bulb should be the only thing that observers can see, all the other components being hidden in a small box on which the bulb is mounted. An LDR is about ⅝in. diameter though even this can be well disguised since the active surface is rather smaller and in any case not all of it has to be exposed; a ¼in. diameter hole should be sufficient; this hole should be close to the bulb and pointing at it. It must, of course, be possible to easily interrupt the light path between the bulb and the LDR in order to "snuff" it.



One Armed Bandit

This circuit is designed to give an approximate simulation of a one armed bandit fruit machine, and is intended for home-entertainment purposes only. The unit has three seven segment LED displays, and when a pushbutton is depressed, all display segments light up. When the button is released, a random number is displayed.

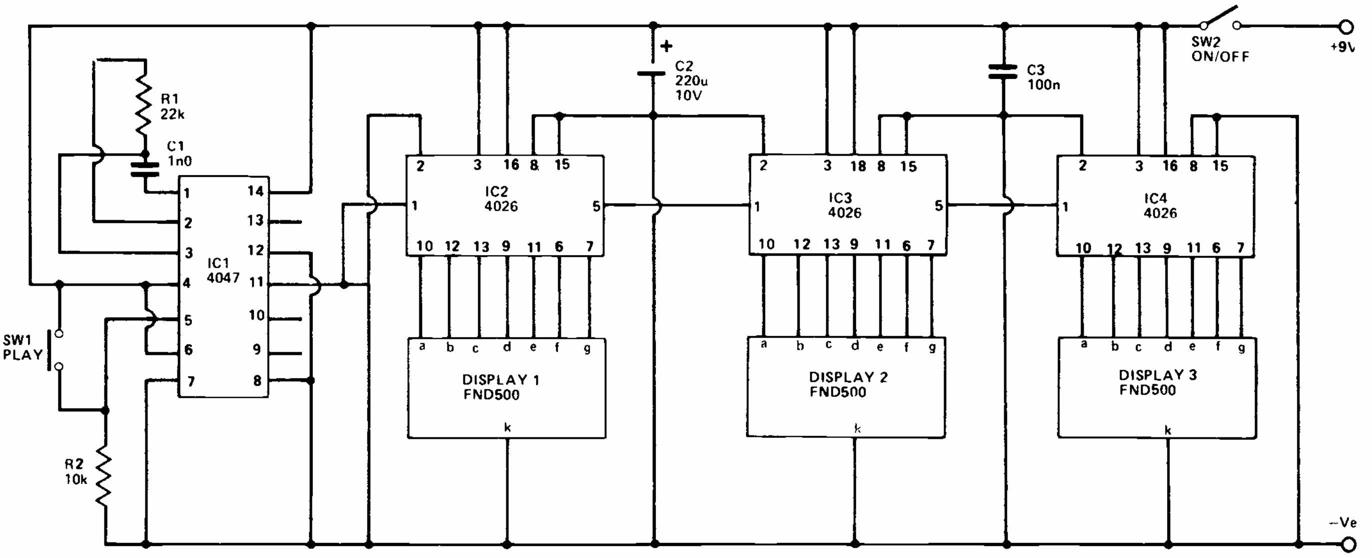
The idea of the game is to obtain a row of three identical numbers in the display, with (say) 1 point being scored for '000', two points for '111', etc., up to 10 points for '999'. The object of the game is to score as many points as possible in an agreed number of attempts, say 25 or 30.

The circuit consists basically of a clock oscillator using IC1 and a three stage counter which uses IC2 to IC4. The 4047 CMOS device is a monostable/astable device which is used here in the true gating astable mode. Under quiescent conditions R2 takes the gating input (pin 5) low, and prevents the circuit from oscillating. Depressing SW1 takes the gating input high, and starts the circuit oscillating at a frequency which is controlled by R1 and C1. The specified values give an operating frequency of about 10kHz; although the exact frequency is unimportant, it does need to be reasonably high.

The display section uses three CMOS 4026 decade counter/seven segment decoders, which can be used to directly drive high efficiency common cathode displays such as the FND500, DL704, etc. The three display circuits are connected in series so that one thousand clock pulses take the display through every number from '000' to '999', and then back to '000' again.

Thus, when SW1 is operated the display cycles through about ten times per second. All the display segments appearing to switch on continuously. When SW1 is released, the display is "frozen" at whatever number it happened to be displaying when IC1 ceased oscillation. There is, of course, no way of operating SW1 to definitely obtain one of the winning numbers, and it is purely a matter of chance whether or not one of these is displayed.

IC1 to IC4 are all CMOS devices, and normal CMOS handling precautions should be observed. The current consumption of the unit can be over 50mA when certain numbers are displayed, and a large battery such as a PP9 should be used.



Quiz Monitor

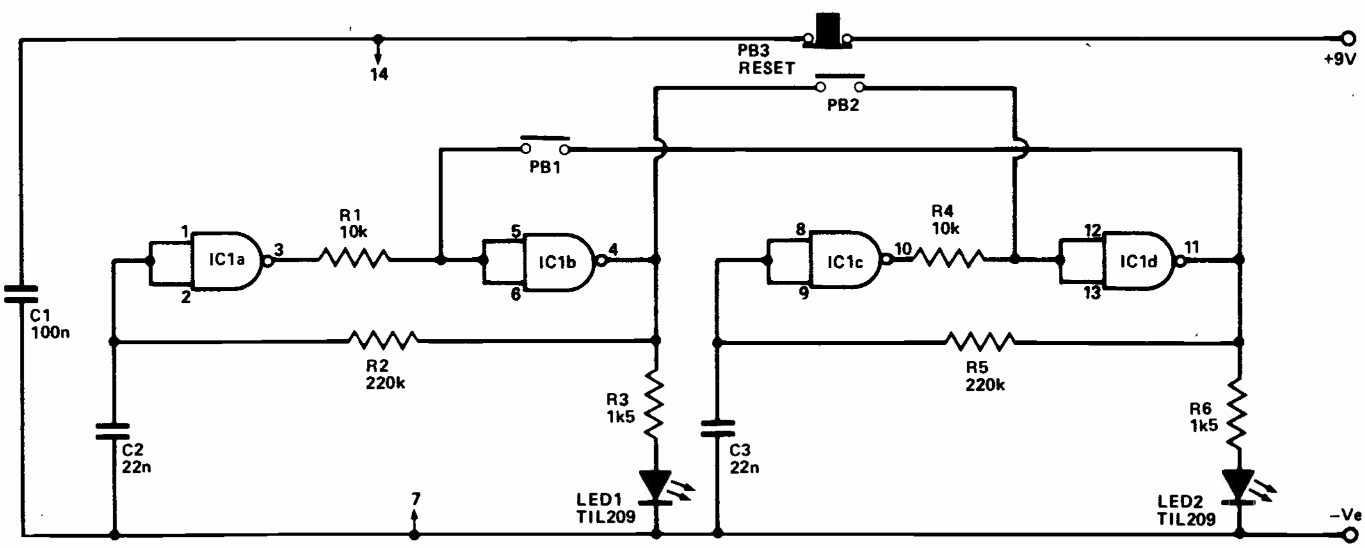
This circuit is useful when playing "snap" or TV quizzes where the first person to have an opportunity of answering the question is the first to operate their push button switch. Operating it causes an indicator light to switch on and prevents the opponent's switch and indicator light from working. Thus, there is no doubt as to which push button switch was operated first.

The circuit is based on a CMOS 4001 quad two input NOR gate, but all of the gates have their two inputs connected together so that they, in fact, operate as simple inverters. The circuit has two identical sections, one for each player, each using two of the gates.

If we consider gates 1 and 2, at switch on C2 is in a discharged state and, therefore, takes the input of gate 1 low. This causes the output of gate 1 to go high, taking the input of gate 2 to the same state due to the coupling through R1. The output of gate 2 then goes low and feedback through R2 holds the input of gate 1 in its original low state, thus latching the circuit in this condition. LED 1 is driven from the output of gate 2 via current limiting resistor R3. At first it will obviously be switched off. So will LED 2 which is the indicator light for the other player. It is driven from an identical arrangement.

If PB1 is activated, the input of gate 2 is taken low, since it will be taken to the low output of gate 4. This sends the output of gate 2 high, the input of gate 1 high (due to the feedback through R2) and the output of gate 1 low. The circuit will hold itself in this state even if PB1 is released. Operating PB2 will have no effect now, since this will merely connect the high output of gate 2 to the high input of gate 4, producing no changes in logic state. Thus the required blocking is obtained. Of course, if PB2 was operated first, LED2 would switch on and PB1/LED1 would be disabled, with the basic circuit action being the same as described above. The operating speed of the circuit is extremely fast and even with a very small gap between the two switches being operated, the unit is capable of determining which was operated first and there is no danger of both LEDs switching on.

The circuit is reset by briefly operating PB3 so that power is removed from the circuit and it starts once again from the beginning when PB3 is released. No on-/off switch is required as the unit has negligible quiescent current.



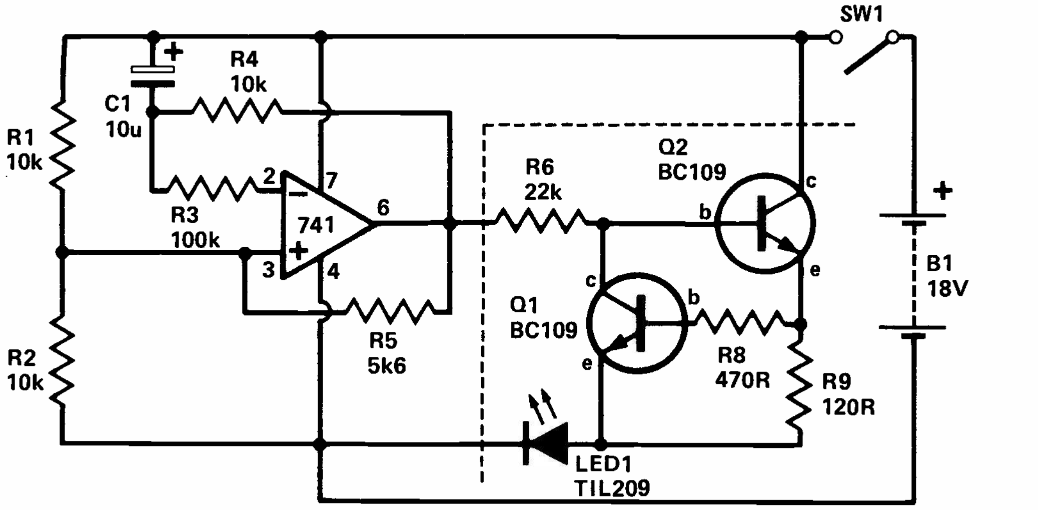
Clocks and Oscillators

Op-Amp Oscillator (1)

This circuit produces a low frequency square wave, using a 741 operational amplifier as a comparator. Indication of the frequency is provided by a light emitting diode (LED), illuminating when the output voltage is high.

By changing the values of C1 (10u) and R4 (10k) the frequency can be altered. Reduction of C1 or R4 increases the frequency; an increase in value reduces the frequency. If the frequency is above approximately 30Hz the LED will appear to be on continuously and the section after the dotted line can be omitted. Then the op-amp output (pin 6) can be fed directly into an audio circuit (via a 10k resistor), and act as a straight-forward signal injector.

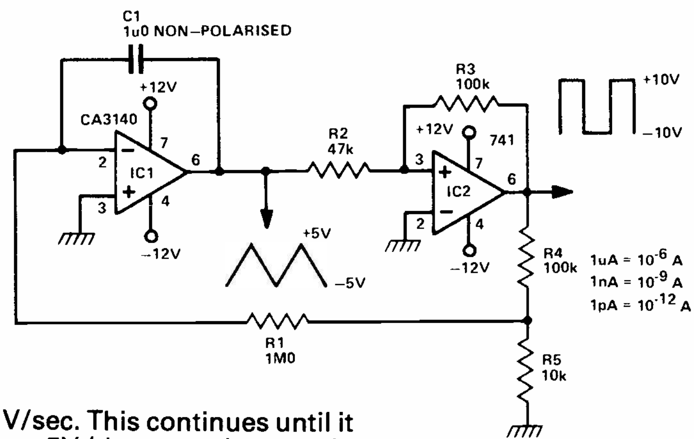
Using the component values shown, the circuit will flash the LED at a rate of approximately 5Hz and if a socket is used, the circuit makes a useful op-amp tester.



To explain how the circuit operates, assume that the output of the 741 is low. The inverting input will be at a point below half supply. Also assume that C1 is not charged. The lower end of C1 will start to fall in voltage (due to the current through R4) until it reaches the voltage at pin 3. At this point the output of the 741 (which is comparing voltages at pins 2 and 3) will go positive. The voltage at pin 3 will rise to the other side of half supply and C1 will repeat, causing the 741 outpout to follow a squarewave.

Low Frequency Oscillator

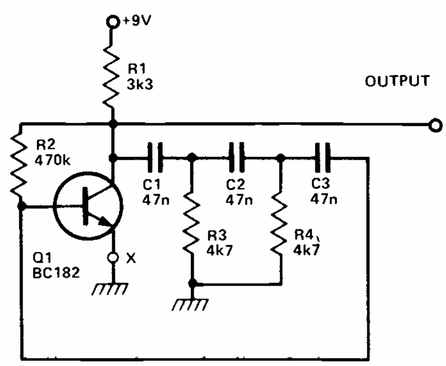
IC1 is a MOSFET op-amp. Thus its input bias current is very low indeed, typically 10pA as compared with 100nA for a 741. This allows very low current designs to be produced. The circuit shows an integrator (IC1) and a Schmitt trigger (IC2). Imagine that the output of the Schmitt is high (+10V). The voltage at the junction of R4, R5 is approximately +1V. This pushes a current of 1uA through R1 which then charges C1. Thus C1 (the output of IC1) ramps down at a rate of I/C which is in this case 1V/sec. When this voltage reaches −5V, the Schmitt trigger flips over into its low state (−10V). Now the current through R1 flows the other way, and the output of IC1 ramps



up at 1V/sec. This continues until it reaches +5V (the upper hysterysis level of the Schmitt). The Schmitt trigger then jumps to its high output and so the period process repeats itself. The circuit produces a square wave (±10V) and a triangle (±5V) output. Using the components shown the period is 20. To get 200 seconds make R1=10MR; to get 2000 seconds make R1+10MR, R5=1R.

Phase Shift Oscillator

A single transistor can be used to make a simple phase shift oscillator. The output is a sinewave with a 'lump' in it, which means that the distortion content is rather high, about 10%. This is not always a problem; quite often when generating audio tones a high harmonic content will make a more interesting sound. The sine wave purity can be increased by putting a variable resistor (25 ohms) in the emitter lead of Q1. The resistor is adjusted so that the circuit is only just oscillating, then the sinewave is relatively pure. However, if the power supply level varies, the oscillation may cease altogether. The operating frequency may be varied by putting a 10kR variable resistor in series with R3, or by changing C1, 2, 3. Making C1, 2, 3 equal to 100nF will halve the operating frequency.



Op-Amp Oscillator (2)

An op amp can be made to oscillate, generating a square-wave output. The circuit is a Schmitt trigger and an integrator rolled into one.

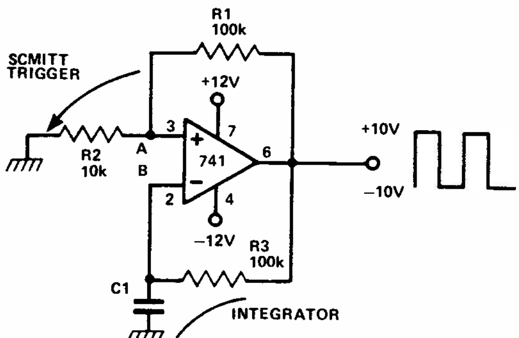
To understand the operation, imagine the output is high; C1 is charged up via R3. The voltage at point 'A' is +0.9 V due to the resistor divider network R1, R2. When the voltage at B exceeds this, the output of the op amp flips into its negative (low) state. C1 is therefore discharged by R3. When the voltage on C1 reaches −0.9 V, the reverse process occurs and the op amp output flips back to its high state. Thus the circuit oscillates producing a square wave going from +10V to −10V.

The frequency of operation can be obtained from the voltage changes on C1. This is the truncated section of an exponential charge/discharge curve, but we shall ignore this and assume that the curve is linear (which it almost is).

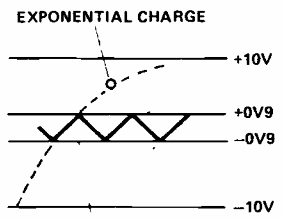
The frequency can be obtained from the formula

$$F = I/\Delta V \times C \text{ Hz}$$

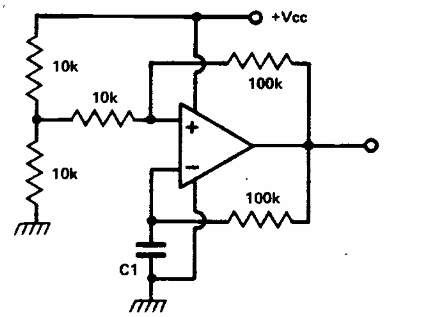
where I is the charging current (approximately 100μA), ΔV is the charge across C1 (3.6 V) and C is the capacitance in Farads.



DUAL POWER SUPPLY OPERATION

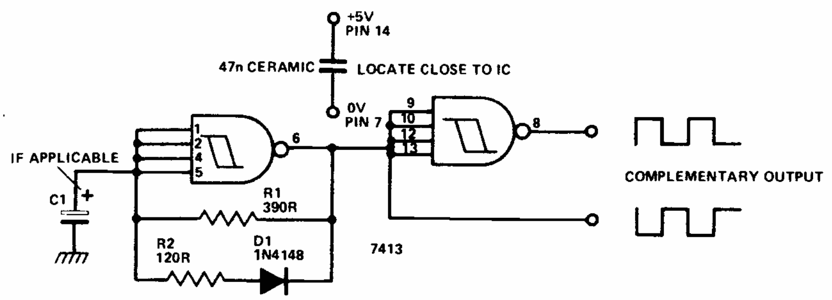


VOLTAGE AT POINT B



SINGLE SUPPLY OPERATION

HF TTL Clock

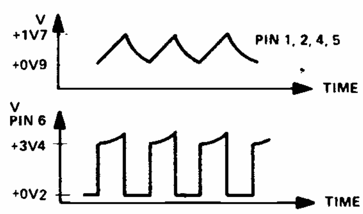


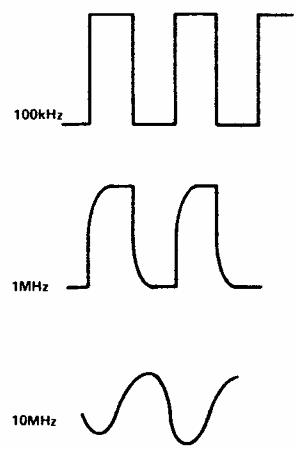
A squarewave oscillator with complementary outputs and a frequency range of 20Hz to 10MHz can be made from one IC, a 7413 which is a TTL dual Schmitt trigger. The oscillator is self starting and runs from a 5V supply, current drain 20 to 30mA. The 7413 is a Schmitt trigger with hysterysis levels (at its input) of +0.9V and +1.7V. That is when the input level exceeds +1.7V the output jumps to a low condition (+0.2V). When the input voltage is lowered it needs to fall below +0.9V before the output jumps back to a high condition (+3.4V).

When the Schmitt trigger is connected as shown the device will oscillate. Imagine the output is high. C1 is charged up via R1. When the voltage on C1 reaches +1.7V, the output falls to +0.2V. C1 is now discharged via R1 in parallel with R2 (D1 is now forward biased) until the voltage on C1 reaches +0.9V. Then the output jumps to a high state and the process repeats itself. The second Schmitt trigger merely inverts the squarewave output. The frequency of operation is given by the formula:

$$F = \frac{2 \times 10^{-3}}{C1}$$

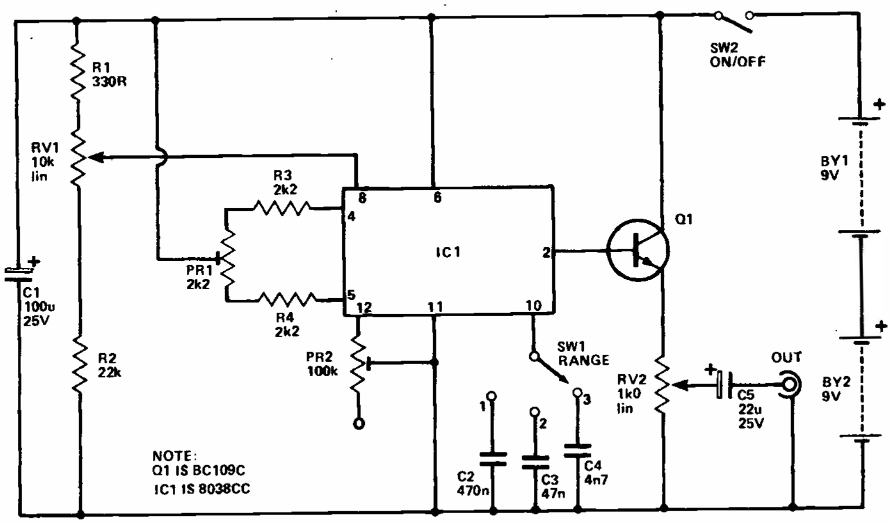
where F is in Hz and C1 is in Farads.





WAVEFORM DEGRADATION WITH FREQUENCY

8038 Signal Generator



Although the 8038CC is not capable of generating an extremely pure sinewave, it is capable of producing an output of high enough quality for general audio testing. The simple circuit shown here covers the audio frequency spectrum in three ranges – less than 20Hz to more than 200Hz; less than 200Hz to more than 2kHz, less than 2kHz to more than 20kHz. The output amplitude is continuously variable up to a maximum of about 550mV rms and is from a low impedance source.

The 8038CC oscillates by first charging a capacitor via a constant current source and then discharging it through another constant current generator, It thus generates a triangular waveform. This is then fed to a trigger circuit to generate a squarewave signal and to a non-linear amplifier which "rounds off" the signal to give a sinewave output of reasonable purity. C2 to C4 give the three ranges. R1, RV1 and R2 form a potential divider circuit, which is used to control the charge and discharge currents of the timing capacitor. RV1 thus acts as the fine frequency control. PR1, R3, R4 balance the charge and discharge currents, so that a symmetrical output is obtained. PR2 is part of the sinewave shaping circuitry and is adjusted for maximum purity.

The sinewave output at pin 2 of IC1 is at a high impedance and is, therefore, coupled to the output via an emitter follower buffer stage using Q1. RV2 is the output level control, and C2 provides DC blocking at the output.

With the unit adjusted for a fairly low frequency output (about 50-200Hz), it should be possible to hear the main fundamental frequency plus the higher frequency harmonic signals. The output can be monitored using a crystal earphone or amplifier/loudspeaker. PR1, 2 are adjusted to minimise the harmonics.

Slide/Tape Synchroniser

With the aid of a tape recorder and a slide tape synchroniser it is possible to obtain programmed slide changing with an automatic projector. By using a synchroniser and a stereo tape deck or recorder it is possible to have music and a commentary recorded on one channel and signals to give automatic slide changes at the appropriate points on the other channel.

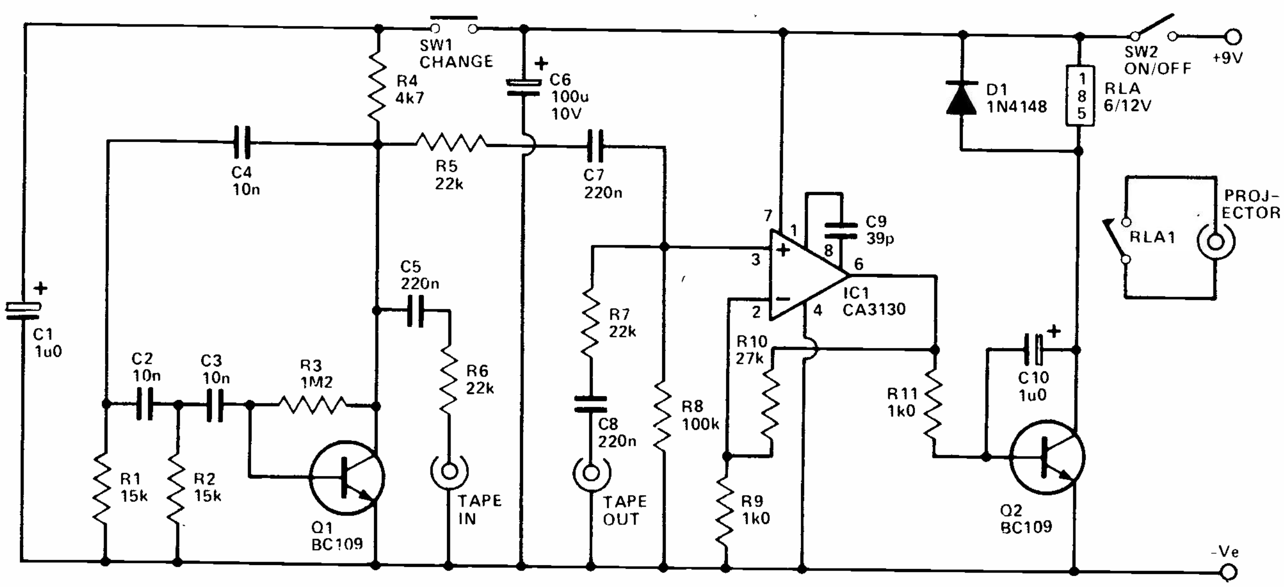
A slide/tape synchroniser has two sections: a tone generator and an electronic switch. The tone generator is used to record short bursts of tone onto the tape at the points where slide changes are required. The electronic switch is fed with the tone burst output of the tape recorder and closes a pair of relay contacts for the duration of each burst. The relay contacts are, of course, used to control the automatic slide change mechanism of the projector. Usually the output of the tone generator is coupled to the input of the electronic switch, so that operating the tone generator causes the relay contacts to close. This is useful when recording a tape. With the projector loaded with slides, the synchroniser connected to the projector, the output of the tone generator fed to one input of the recorder and the music/commentary signal ready to be fed to the other input, the tape is inserted. Then the music and commentary are recorded and the tone generator is operated at the appropriate times so that the slides are changed and the tone bursts are recorded. If the tape is then rewound, the slide magazine is brought back to its starting point and the tone burst output of the tape recorder is fed to the input of the electronic switch, replaying the tape should give the slide show with accompanying sound track and automatic slide changing. The operator only has to start the tape at the beginning of the show and stop it at the end.

A similar technique is used when using the unit as a programmed slide timer, the only difference being that there is no soundtrack to bother with.

The tone generator uses Q1 in a straightforward phase shift oscillator operating at about 500Hz, although the exact operating frequency is not of great importance. The output from the collector of Q1 is coupled to the tape recorder by DC blocking capacitor C5 and resistor R6. The latter attenuates the output. R6 also ensures that the oscillator cannot be so heavily loaded that it ceases functioning. SW1 is a non-locking, push-to-make switch. It is pressed briefly to connect the supply to the tone generator and produce the tone bursts.

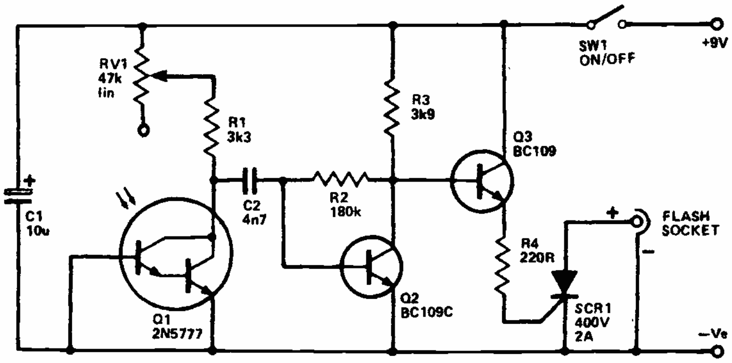
The electronic switch is based on operational amplifier IC1, which is used in the non-inverting mode. Its voltage gain is set at about 28 by R9, 10, and R8 biases the non-inverting input to the negative supply rail. R5, 7 form a simple passive mixer at the input of IC⅛, so that it can be fed from either the tone generator or from the output of the tape recorder without the need for any change-over switching. The output of IC1 is used to drive common emitter amplifier Q2, which has the relay coil and protective diode D1 as its collector load. Normally IC1's output is low and Q2 is cut off, but in the presence of an input tone the output of IC1 goes strongly positive on positive going half cycles. C10 integrates these pulses so that Q2 is continuously switched on in the presence of an input tone and the relay is energised, the relay contacts close and operate the slide change mechanism of the projector.

The current consumption of the unit is only about 500μA, but rises to around 40mA during the brief periods when the relay is activated. The relay can be any type having a 6/12V coil with a resistance of about 185R or more, provided it has at least one set of normally open contacts of adequate rating.



Photography

Flash Slave Unit (1)



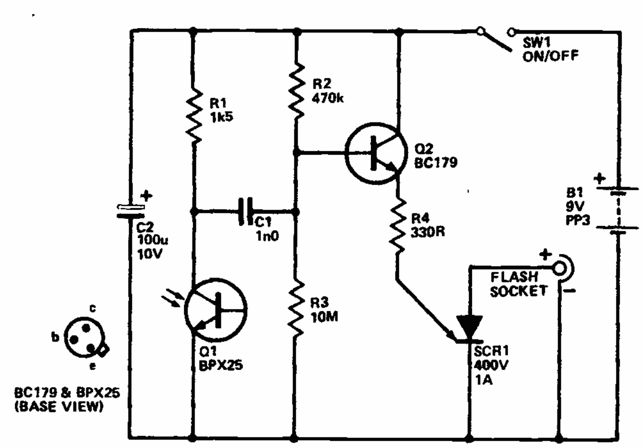
The photocell used in this circuit is a photo-Darlington transistor. This gives a fairly fast operating speed and high sensitivity. In fact the sensitivity is rather too high, making it likely that the cell would saturate in only moderate light. Its base terminal is, therefore, connected to the negative supply rail to give a suitable reduction in sensitivity. R1 and RV1 form the collector load for photocell Q1 and RV1 acts as a sensitivity control. With RV1 at a low resistance, the increase in the current passed by Q1 when it picks up the pulse of light from the primary flashgun will produce a fairly small voltage spike across the load resistance. With RV1 set at a high resistance, a similar current pulse would produce a much larger voltage spike across the load and high sensitivity is obtained.

One problem with equipment of this type is that under bright conditions the photocell can saturate, preventing the circuit from functioning. When used indoors, saturation is unlikely to occur even with RV1 set for maximum sensitivity. The sensitivity of the unit should be so high that it will trigger reliably even if the primary flashgun and Q1 are aimed in opposite directions. When used outside in bright conditions it would be advisable to back off RV1 and the aim of Q1 and the flashgun will inevitably be more critical (there will probably be less reflected light to trigger the unit in addition to the reduction in sensitivity).

C2 couples the output from Q1's collector to the input of a common emitter amplifier, Q2. This is biased by R2 so that there is a quiescent collector voltage of only about 1V. Q3 is an emitter follower buffer stage which is used to drive the gate of SCR1 from Q2's collector. The quiescent voltage at Q3's emitter is insuffient to activate the thyristor, but when Q2 receives the negative voltage spike from Q1 it switches off and the emitter potential of Q3 rises to a high enough level to trigger SCR1 and fire the second flashgun. R4 is a current limiting resistor which prevents Q3 from passing an excessive current.

The current consumption of the circuit is about 2mA. Note that the flash lead must be connected to SCR1 with the correct polarity or the unit will not operate.

Flash Slave Unit (2)



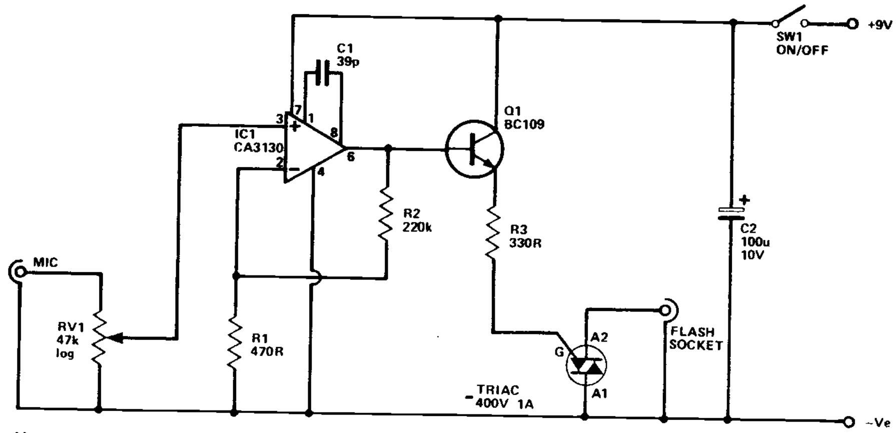
The problem of harsh and unnatural shadows that occurs when taking photographs using a single flash can be overcome by the use of a second flashgun. A flash slave unit is then very useful as it enables the additional flashgun to be triggered by the light from the main gun, reducing the amount of interconnecting leads required.

The flash slave circuit shown here is extremely sensitive and when used indoors it always triggers reliably regardless of where the main flashgun and the slave unit are positioned. If used out of doors there is likely to be little reflected light to trigger the slave unit and it may be necessary to have the photocell aimed towards the main flashgun in order to obtain reliable operation. The circuit has excellent immunity to saturation by strong ambient lighting.

The photocell is a BPX 25 or similar silicon photo-transistor. The collector to emitter resistance of this component varies over a wide range from darkness to high lighting levels, with the latter giving the lowest resistance. The photocell is connected as part of a potential divider across the supply rails and when the light from the flash unit is received, a negative output pulse is produced. This pulse is coupled by way of C1 to the input of a common emitter amplifier based on Q2. C1 has been given a low value so that it only efficiently couples the fast pulse caused by the flashgun, and slow signals such as those caused by shadows crossing the photocell are blocked.

R2 and R3 forward bias Q2, but by an amount which is too small to produce a significant collector current. However, the negative input from the photocell causes Q2 to conduct heavily and trigger SCR1 gate via current limit resistor R4. SCR1 then briefly switches on and fires the flash unit. The circuit operates extremely rapidly and there is no significant delay between the firing of the two flashguns. The unit has a current consumption of about 20 to 50μA in normal ambient lighting conditions, falling to less than a microamp if it is stored in darkness and it is therefore quite feasible to omit the on/off switch SW1.

Sound Triggered Flash (1)



By triggering an electronic flashgun using a sound operated switch, photographs of such things as a balloon bursting, the cork leaving a champagne bottle and objects splashing into water can be taken. Since electronic flashguns normally give an effective shutter speed of around a 1000th of a second, a "frozen" action photograph is obtained.

The photograph must be taken under fairly dark conditions so that the ambient light does not give an exposure if the camera's shutter is set to "B" and opened.

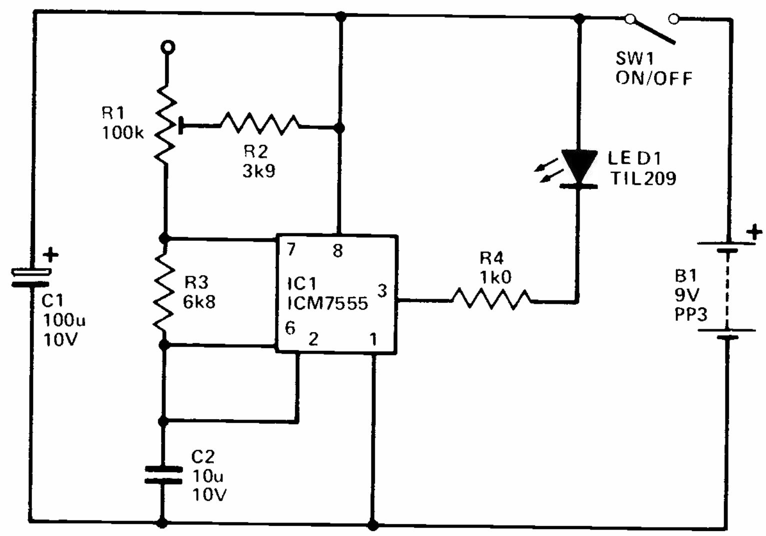
The circuit is based on operational amplifier IC1 which is used in the non-inverting amplifier mode. R1, 2 are a negative feedback network which set the gain of the unit at about 500. RV1 (sensitivity) biases the non-inverting input to the negative supply rail. Ideally the input should be fed from a crystal or high impedance dynamic microphone, but the unit will work quite well using a low impedance dynamic microphone or even a high impedance speaker as the signal source.

Q1 is used as a discrete emitter follower output stage which provides the relatively high trigger current required by the triac. R3 is a curent limiting resistor. Under quiescent conditions the output of IC1 will be at virtually negative supply potential, and the triac, therefore, receives no gate current. When a signal is received by the microphone, positive going signals are amplified by IC1 to give an output that is a few volts positive. The triac then rceives a strong gate bias, causing it to trigger and give a low resistance across its A1 and A2 terminals. These terminals connect to the flashlead via a suitable socket (or flash extension lead with the unwanted plug removed) and the flashgun is, therefore, fired. The circuit operates almost instantly, giving very little delay between the commencement of the sound and the flashgun being triggered. Sometimes more interesting photographs can be obtained by introducing a small delay.

This can be achieved by moving the microphone a metre or two away from the object(s) being photographed.

The current consumption of the unit is approximately 4mA. It is advisable not to advance RV1 much more than is absolutely necessary in order to give reliable triggering, as frequent spurious operations of the unit could otherwise result.

Simple Photo Timer



Although this timing device may seem to be rather unsophisticated, it is a handy little gadget for timing darkroom exposures, or time exposures, or time exposures made on a camera with the shutter set to the "B" position. The unit simply flashes a LED indicator briefly at 1 second intervals. Adequate accuracy for normal requirements can be obtained in this way.

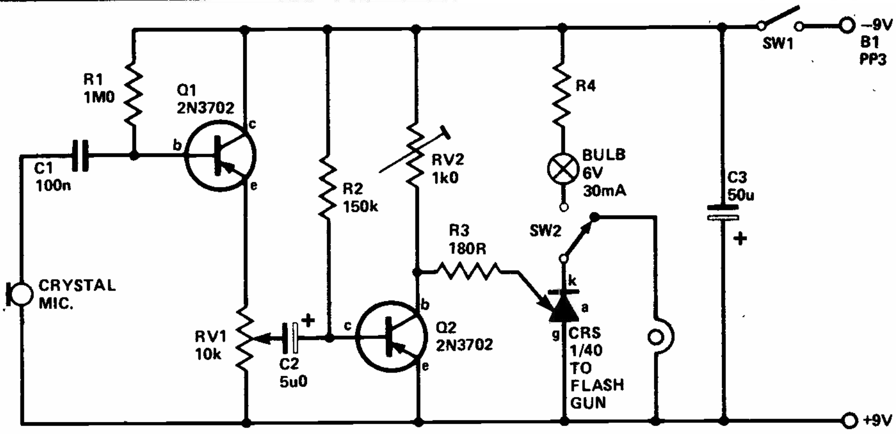
The circuit is based on the CMOS version of the well known 555 timer device. The average current consumption of the unit is less than 1mA, giving an extremely long battery life.

The CMOS version of the 555 operates in the same manner as the ordinary version, with timing capacitor C2 first charging up to ⅔ V+ by way of the timing resistors R1, R2, R3. The device is then triggered into the discharge mode, resulting in C2 being discharged through R4 to a potential of ⅓ V+ whereupon the circuit reverts to its original state with C2 charging up once again. Continuous oscillation thus results. The frequency of operation is adjusted to 1 HZ by adjusting R1, and in practice this is adjusted by trial and error to obtain (say) 60 flashes in a one minute period. Longer calibration periods can be used if better accuracy is required.

The output of IC1 assumes the high state while C2 is charging, and the low state while it is discharging. As C2 charges via R1, R2 and R3, but only discharges through R4, the discharge time is therefore much shorter than the charge time. By connecting LED indicator D1 and its current limiting resistor R4 between the output of the IC1 and the positive supply, brief flashes are obtained.

Photography

Sound Triggered Flash (2)



The circuit shown is completely solid state and instead of a relay, a SCR is employed. This is cheaper and, for this function, just as good.

The first stage of the circuit is an impedance convertor. A crystal make is used. Normally these have rather poor quality but in this circuit we are not too interested in quality; we are only using it as a device for converting sound into an electrical pulse. Q1 is connected as a common collector stage; this has very high input impedance to correctly match the high impedance of the crystal microphone. The potentiometer RV1 is the emitter resistor and the sounds produced appear across this at a workable impedance. The output is fed to the conventional common emitter amplifier, Q2 with RV2, a preset pot, as the collector load. The collector of this transistor is connected to the gate of the SCR via resistor R3. For setting up, the SCR is connected to a bulb.

When a sound is produced it is amplified by Q1 and Q2 and causes Q2 to draw rather more current at the peak of the sound. This reduces the voltage at the collector of the transistor, and this is fed to the gate of the SCR. At the correct setting of RV1 this will cause the SCR to switch on and light will pass through the bulb. The bulb can be a 9V type but as these are hard to come by it can just as well be a 6V type with a 33 ohm resistor in series.

The bulb is used only for setting up. To continually trigger the flash gun in order to find the correct settings will be wasteful, especially as the flash tube has a limited life. Once the correct settings have been found, SW2 can be made and the SCR applied across the flash gun terminals. There are two variables in the circuit, RV1 and RV2. RV2 will normally require setting once only.

With the slider of RV1 about a quarter the way up the track from the positive line, RV2 should be set so that the SCR just triggers on the loudest sound that can be made near the microphone. When this is done RV1 should give control over a wide range of sounds and acts as the sensitivity control.

The circuit should be tested, to obtain the correct level setting of RV1, before every shot is taken with the test bulb in circuit. Once the correct settings are obtained the switch can be made to the flash gun, having first made sure that the SCR is not on at that point. The SCR will stay switched on until the supply voltage is removed and it is necessary to switch off using SW1 before switching over.

Precision Half-Wave Rectifier

By putting a diode in the feedback loop of an op amp, a precision rectifier can be constructed. Normally a diode will need to be forward biased by 0V6 before it will start conducting, so if you want to half wave recify a low sine-wave (say 0V1 peak-to-peak) it is almost impossible to do it with just a diode.

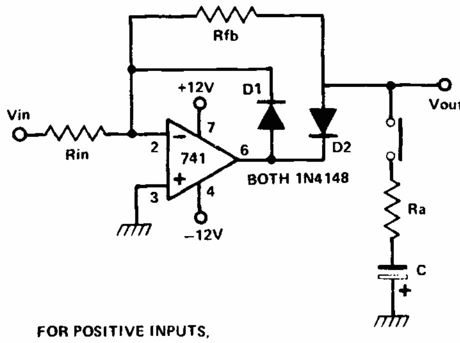
However, by using an op amp it is possible to get this diode to drop to below 1mV in most applications. Referring to the circuit, imagine Vin goes positive; then the output of the op amp will swing negative to such an extent that D2 will be properly biased and will draw current through the feedback resistor Rfb. In fact the op amp adjusts its output so that the voltage at pin 2 is virtually at 0V (virtual earth). Thus the output voltage:

Vout=Vin x Rfb/Rin

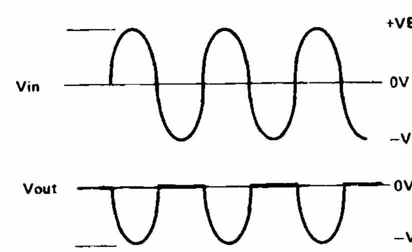
which is just like a normal op amp.

The diode D2 doesn't seem to have affected things and this is true for positive inputs even as low as a few millivolts. When Vin goes negative, the output of the op amp swings positive, D1 conducts and maintains a virtual earth condition and D2 is reverse biased. So, now the output is just Rfb connected to effectively 0V. What this means is that there is only an output voltage (negative) for positive going inputs; when the input goes negative, the output is zero. That is, the input signal has been precisely half wave rectified.

Now, if the Ra, C network is connected to Vout, the half-wave rectifier can be turned into a negative envelope follower. When Vout goes negative, C is charged via Ra, Vout is unaffected wilst C is being charged. When Vout returns towards 0V, C discharges through Ra and Rfb. If C and Ra are correctly selected then a contour of the envelope of the signal will be produced at Vout. For an envelope attack time of 1 millisecond (1kMz), with an envelope release time of 100 milliseconds, make Rin and Rfb 100k, Ra 1k and C 1uf.



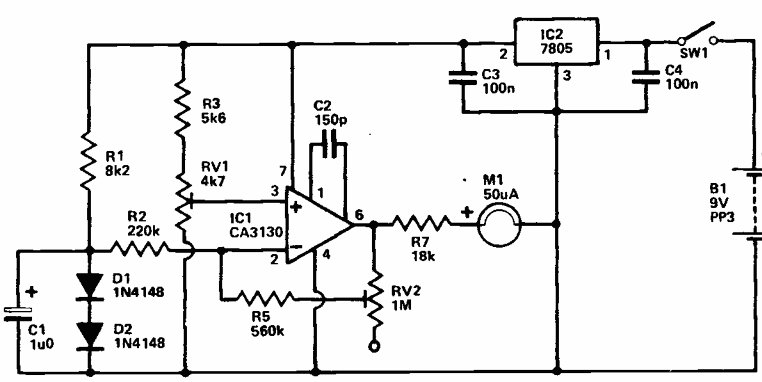
FOR POSITIVE INPUTS,
VOLTAGE GAIN = − Vout/Vin = Rfb/Rin



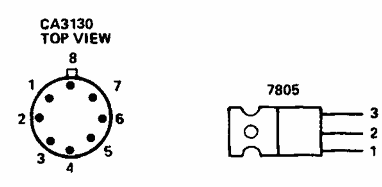
Electric Thermometer

This thermometer covers the range 0 to 50 degrees Centigrade with a linear scale so that the temperature can be read directly from a 50uA meter. A range of 0-100 degrees Centigrade can be obtained by substituting a 100uA meter.

The unit uses silicon diodes D1 and D2 as the temperature sensors and these would normally be mounted in some form of probe which can be positioned many metres away from the other circuitry, if necessary. C1 filters out any noise picked up in the connecting cable. D1 and D2 are given a small forward bias by R1, so small that there is no significant self heating of the diodes. The voltage produced across the diodes is nominally 1V2 but it actually varies by about 2mV per degree C per diode, or about 4mV across both diodes. This voltage is fed to the input of an op amp inverting amplifier, IC1. With the probe at 0 degrees C (which can be achieved by immersing the probe in ice) RV1 is adjusted for the highest voltage at IC1's non-inverting input that gives zero output voltage. This compensates for the quiescent voltage across the diodes, and gives zero reading on the 1V FSD voltmeter circuit connected across the output of the amplifier.



If the diodes are heated to 50 degrees C the voltage across them will fall by approximately 200mV, this is amplified by a factor of 5 by the amplifier to give about 1V at its output, roughly full scale deflection of the meter. In practice RV2 is used to adjust the gain of the amplifier so that precisely full scale deflection is produced.



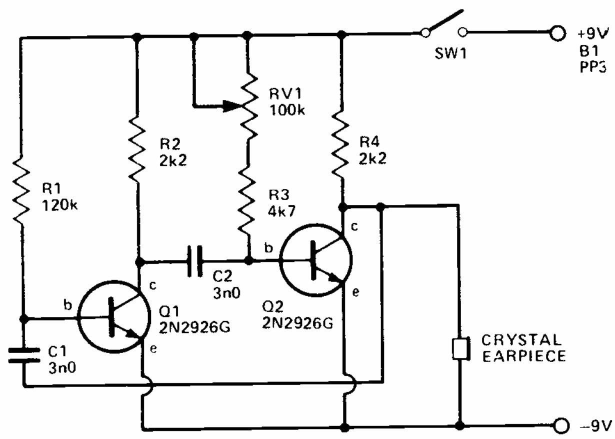
Of course RV2 can be given the correct setting with the probe at any known temperature which corresponds to a reasonably substantial meter deflection.

The circuit requires a very stable supply of about 5V, obtained from a 9V battery using a 5V monolithic regulator (IC2). C3 and C4 should be mounted close to IC2 in order to prevent instability.

Miscellaneous

Insect Repellant

It seems that mosquitoes and other nasty insects only mate at certain times and except for these times the two sexes are most unfriendly, in fact they stay well away from each other. It has also been reliably established that it is only the female of the species that actually bites. The third fact that we need to know is that the male mosquito (and this applies to other bugs as well) beats its wings at a slightly different rate than the female – this is one way that they identify each other. From these gems of information it will be seen that if one electronically simulates the sound of a male mosquito, the females will steer well clear.



The circuit shown is a simple audio oscillator whose frequency of operation can be varied over a wide range, in fact from about 500Hz to 10kHz and this will take in the range of all the common bugs. The circuit is a straightforward multivibrator with RV1 altering the audio frequency. This produces a square wave which is applied across the small crystal earpiece connected between the collector of Q2 and the negative line. Crystal earpieces have a very high impedance and it will not affect the operation of the circuit. Almost any transistors can be used in this simple circuit but if PNP types are used the battery supply should be reversed. The values of the capacitors are not too critical either and if others are used and it is found that the frequency range is not adequate, R1 can be altered to bring it back to the right sort of range. The current consumption is low, 2-2mA and varies slightly with the frequency, but a PP3 battery will last for a while; after all, the unit will have to be left on for long periods. None of the components need be large and the unit can be built in a small box to fit into a jacket pocket with the components arranged so that the earpiece is external.

Adjusting for the right frequency is a matter of trial and error.

1. SMALL SIGNAL AND PRE-AMPLIFIERS

Preamplifiers are used to convert low level signals into higher level signals with low output impedances. The primary design considerations for preamplifiers are input impedance, signal gain, equalisation, noise performance and distortion. It is now possible to buy integrated circuits that will perform virtually all common preamplifier functions. Only rarely, (and with a lot of skill), will one be able to improve upon a monolithic 'best' solution with a discreet component design. However, it is necessary to fully understand the operation of the IC if optimum results are to be obtained.

The Op Amp Building Block

Most IC preamplifiers operate as op amps. They have a single or differential input, a low impedance output and a large frequency dependent voltage gain, (Fig.1). Note that the

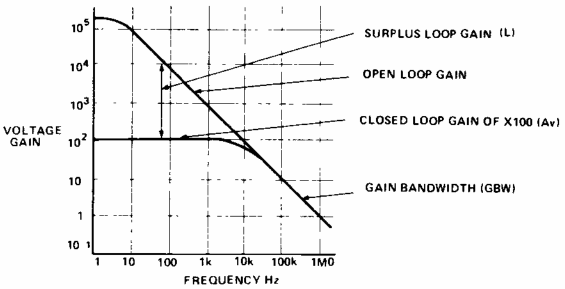
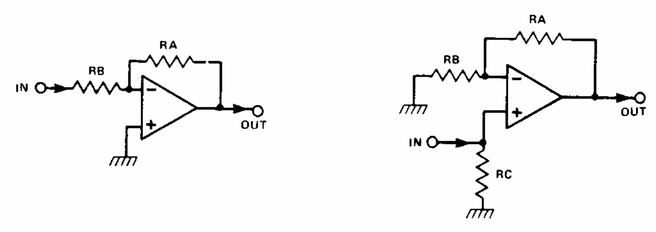


Fig.1. Typical frequency response for the 741 op amp.

open loop voltage gain rolls off at –6 dB per octave. By applying feedback around the op amp (closed loop), the gain is stabilised and is held constant by the resistor ratio until the device runs out of bandwidth. The difference between the open and closed loop gain is known as the surplus loop gain. This surplus loop gain is the negative feedback that is used to iron out non-linearities in the op amp. As the size of the surplus loop gain decreases with increasing frequency, the distortion and output impedance increase. Fig.2 compares the performance of inverting versus non-inverting configurations.



INVERTING MODE

(CLOSED LOOP VOLTAGE GAIN) $A_v = \frac{RA}{RB}$

(INPUT IMPEDANCE) $Z_{IN} = RB$

(OUTPUT IMPEDANCE) $Z_o = \frac{R_o}{1+L}$

CLOSED LOOP OUTPUT THD $= \frac{THD}{1+L}$

CLOSED LOOP BANDWIDTH $= \frac{GBW}{A_v+1}$

NON INVERTING MODE

$A_v = \frac{RA + RB}{RB}$

$Z_{IN} = RC$ *SO LONG AS RC IS SMALL COMPARED TO THE INTERNAL INPUT RESISTANCE OF THE OP AMP

$Z_o = \frac{R_o}{1+L}$

CLOSED LOOP OUTPUT THD $= \frac{THD}{1+L}$

CLOSED LOOP BANDWIDTH $= \frac{GBW}{A_v}$

L = SURPLUS LOOP GAIN
Ro = OPEN LOOP OUTPUT IMPEDANCE
THD = OPEN LOOP THD
GBW = GAIN BANDWIOTH – FREQUENCY AT WHICH OPEN LOOP GAIN IS UNITY

Fig.2. A comparison of the performance of inverting versus non-inverting configurations.

Note that for unity gain the non-inverting mode has twice the closed loop bandwidth of the inverting mode. Also note that the output impedance rises as the surplus loop gain decreases. This can cause a sharp increase in distortion when the op amp is driving a low impedance load at high frequency.

Design Example

Q. Design an amplifier with a gain of 60 dB, a closed loop bandwidth of at least 20 kHz and input impedance of 10 k.

A. First, try the inverting mode as shown in Fig.2.

Zin = 10 k = RB.

$$A_v = 60\,dB = X1000 = \frac{RA}{RB}$$

Therefore, RA = 10 k x 1000 = 10 M

If we use a 741 then the input offset voltage will be multiplied by the closed loop gain. The offset is typically ± 1 to 5 mV. Therefore, the output offset will be ± 1 to 5 V! It is possible to null out the input offset (Fig.3a) with a preset. This circuit will not be very satisfactory.

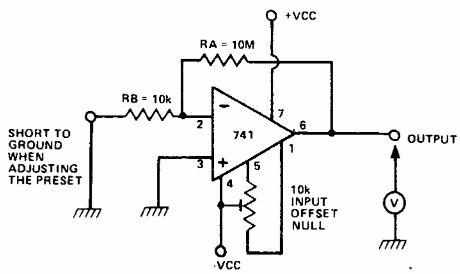


Fig.3a. The input offset can be nulled with a preset.

Its DC output offset will probably drift with temperature and time. If stable high DC gains are required then a 741 should not be used. A high performance instrument op amp should be selected in its place. Another problem in using the 741 is its 1 MHz gain bandwidth product. A closed loop gain of 60 dB will result in a closed loop bandwidth of 1 kHz, and not the 20 kHz needed. An op amp capable of giving 60 dB of gain at 20 kHz would need a gain bandwidth product of 20 MHz.

Although there are some op amps with this performance, they are generally difficult to stabilise and are relatively expensive. A cheap solution is to use two 741s, both with gains of 30 dB (x 33), (Fig.3b). The 741 has a bandwidth of 30 kHz at

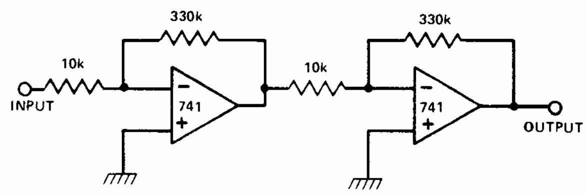


Fig.3b. A cheaper way to achieve higher gain at, say, 20 kHz without incurring a huge bandwidth is to use two 741s.

a closed loop gain of 30 dB. The DC offset still remains a problem, but it could be removed with a nulling preset on the first op amp. However, if the amplifier is to be used for audio then a DC response is not needed and so AC coupling can be used to reduce the final DC offset, Fig.3c. Yet another problem still exists. With the input short circuited the circuit would probably produce about 10 mV of noise at its output. The subject of noise will be dealt with later; suffice it to say that the 741 is not a low noise device. Low noise operation can be obtained by using one of the low noise op amps that are now available.

FOR 20Hz OPERATION

$$\frac{1}{2\pi\, C \times 10k} = 20Hz$$

$$\text{THEREFORE } C = \frac{1}{2\pi \times 10k \times 20} = 0.8u$$

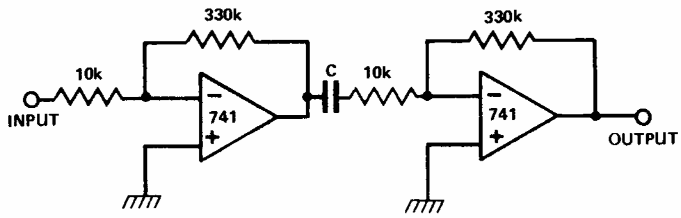


Fig.3c. AC coupling can be used to produce the final offset in audio applications.

Input Offset Voltage And Bias Current

When using op amps as DC amplifiers there are several sources of errors, Fig.4. The input to an op amp is usually an NPN differential pair. To make the device operate, a base current must be supplied (input bias current). Also, to balance the op amp the current through both transistors must be

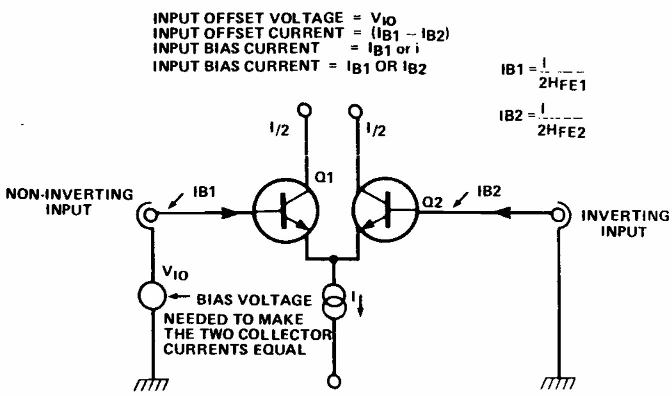


Fig.4. Typical differential op amp input.

equal. The transistor pair is 'matched' for parameters such as Hfe and Vbe versus I_{CE}. However, small differences are caused by the manufacturing process, resulting in the base currents being different (input offset current) and also the base-emitter voltage parameter (input offset voltage). The input offset voltage is multiplied by the closed loop DC gain of the op amp (Fig.5) and the input bias current sets up a DC offset across any resistors it flows through (Fig.6). Typical parameters for the 741 op amp are; 2 mV (V_{IO}), 80 nA (IB) and 20 nA (IB1−IB2). These parameters vary from device to device and from manufacturer to manufacturer. The purpose of successful design is to produce circuits that are insensitive to these variations. Generally, for audio designs, the DC offsets may be eliminated by AC coupling and other methods (Fig.7). Note that the DC output offset may be reduced by inserting a resistor from the non-inverting input to ground. Without that resistor the DC offset may well have been ± 26 mV (IB2 x RA) as opposed to ± 6.6 mV ((IB1−IB2)xRA). A DC voltage on the output would generate a disturbing crackle when the level pot is adjusted.

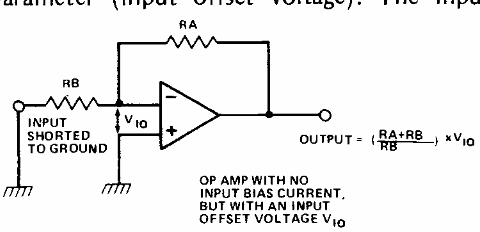


Fig.5. The effect of V_{IO}.

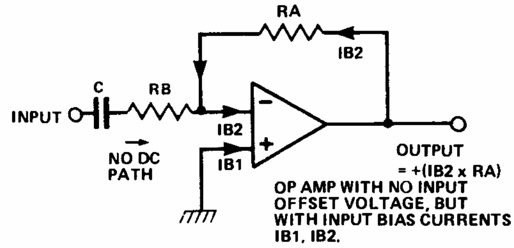


Fig.6. The effect of I_B.

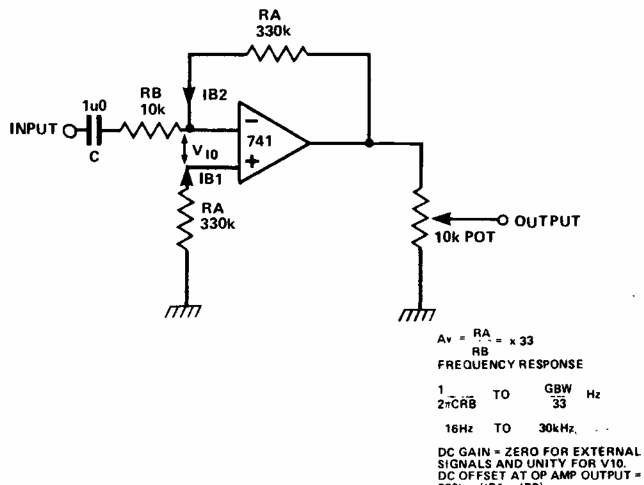


Fig.7. An amplifier with a gain of 30 dB.

Voltage Swing, Power, & Bandwidth

The voltage swing at the output of an op amp is limited in many ways, Fig.8. Most op amps can only swing within a few volts of either supply rail. Also, the speed at which the output voltage can move is limited by the slew rate of the device, which is typically 0V5 per microsecond for a 741. This is the limiting factor in designing amplifiers for high level large signal voltages.

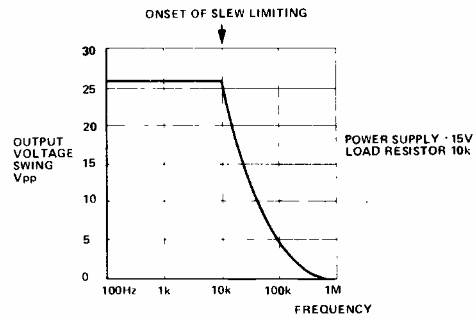


Fig. 8. Typical power bandwidth for a 741 op amp.

Design Example

Q. Design an amplifier with a gain of 4 to amplify a 50 kHz 5 V squarewave.

A. First try a 741, Fig.9. The required output swing is ± 10 V. The 741 can only move at 0V5/uS and so in the half period of the square wave, it can only move 5 V. If the 741 is replaced with a faster device, the TL081 (13 V/uS) then a 'square' wave is produced. The TL081 would take approximately 1.5 uS to travel the 20 V distance, thus generating a

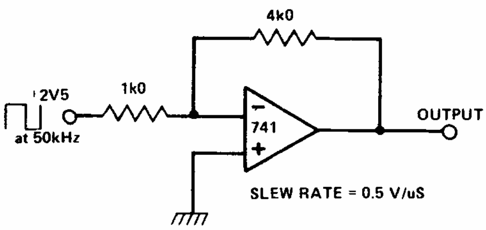


Fig.9. The difference in slew rate between a 741 and a TL081.

rise and fall time of 1.5 uS. Note, that if the resistors were increased in value to say 250 k and 1M0, then the squarewave output would ring. This is because the feedback would have to charge-up the stray capacitance at the inverting input, thus generating a lag between the output and the feedback, which in turn would generate an overshoot.

Noise

Noise is always a problem in electronics. The presence of noise degrades the quality of the signal we are interested in. Everytime we amplify, process, transmit, record or replay a signal, noise is introduced thus worsening the signal to noise ratio. Some common signal to noise ratios are shown below.

Telephone 20 to 40 dB
Cheap cassette player 30 dB
Good tape recorder 60 dB
Professional studio equipment 80 dB

The calculations of noise produced by electronics is complex, but with a few short cuts it is possible to get some useable calculations.

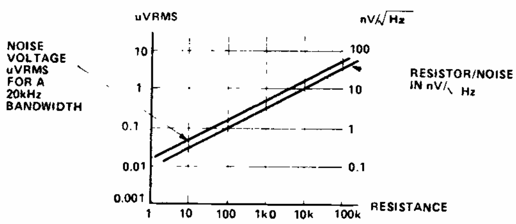


Fig.10. Thermal noise generated by a resistor.

All resistors generate noise due to thermal agitation (Fig.10). Noise is also generated when a voltage is applied to them. Manufacturers generally express this latter noise in uV/V typically 0.1uV/V for metal film devices. For most purposes resistor noise is not a dominating noise source although low level amplifiers perform slightly better with metal film devices. Keeping the resistor values low, helps to obtain low noise operation.

An op amp has several sources of noise generation, Fig.11. there are two noise current generators which both generate noise by flowing through the resistors in the circuit. The resistors themselves generate noise and there is an input voltage noise generator. The total output voltage Eo is given by

$$E_o = \left(\frac{RA+RB}{RB}\right) \times \sqrt{(\text{noise voltage source})^2 + (\text{noise from In+Source})^2 + (\text{noise from In-source})^2 + (\text{noise from RC}) + (\text{noise from RA//RB}}$$

$nV/\sqrt{Hz}$

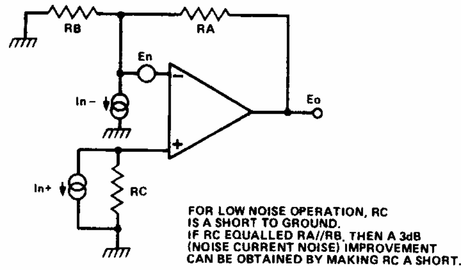


Fig.11. Model of op amp noise generation.

The noise performance curves for a low noise op amp, the SE5534, are shown in Fig.12a,b,c. Graph a shows the input noise voltage density, En (ie total RMS noise in a 1 Hz bandwidth at that particular frequency), as a function of frequency. To convert this input noise voltage, $(4nV/\sqrt{Hz})$ into an equivalent input noise generator, we must define the bandwidth of interest. As the noise spectrum is relatively flat above 100 Hz then we can say,

Fig.12a.The SE5534 low noise op amp input noise voltage density.

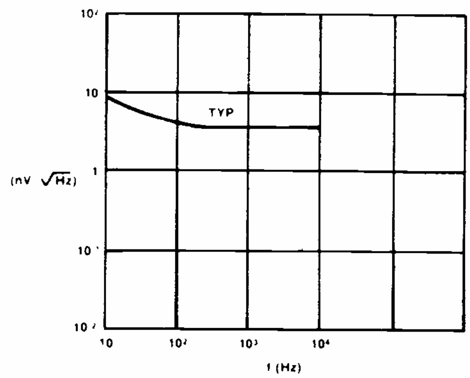
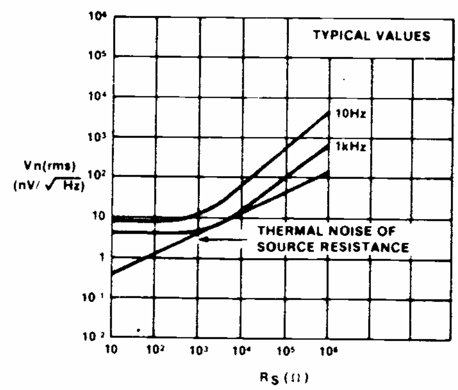


Fig.12b. The variation in total input noise density with source resistance for two frequencies.



Experimenter's Design Notes

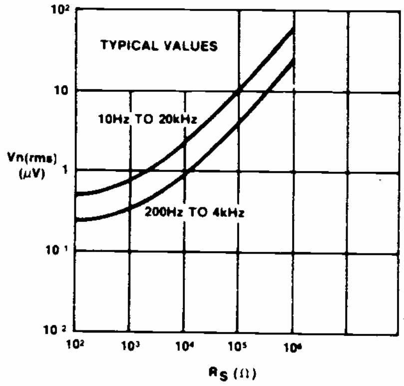


Fig.12c. The equivalent input noise voltage for two bandwidths as a function of frequency.

The equivalent input noise voltage $Ein = \sqrt{(En^2 x bandwidth^2)}$
For 20 kHz bandwidth

$$Ein = \sqrt{(En^2 x 20{,}000^2)}$$

$$Ein = (En \times 141) \text{ nV}_{RMS}$$

$$\text{But } En = 4 nV\sqrt{Hz}$$

$$\text{Therefore } Ein = 4 \times 141 = 0.564 \text{ uV}_{RMS} \text{ (20 kHz)}$$

Design Example

Q. Calculate the output noise in a 20 kHz bandwidth for the circuit in Fig.13. Assume the voltage and current densities have a flat spectrum (which is not too far from the truth!).

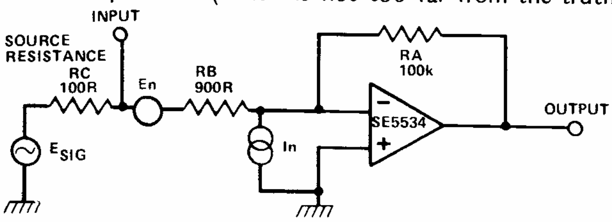


Fig.13. A low noise amplifier.

A. Calculate the individual noise sources.
Resistor Noise
Effective resistor is RA//(RB+RC) 1k0
Therefore thermal noise (from Fig.13b) = 3 nV/$\sqrt{Hz}$
Noise Voltage
a = 4 nV/$\sqrt{Hz}$
Noise Current
c = 0.5pA/$\sqrt{Hz}$
which sets up a noise voltage through (RB + RC). This noise voltage is 0.5pA x 1k0 = 0.5 nV/$\sqrt{Hz}$
Therefore the total noise voltage

$$Eo = \left(\frac{RA+RB+RC}{RA+RB}\right) \times \sqrt{(4)^2 + (0.5)^2 + (3)^2}$$

$$= 101 \times \sqrt{16 + 0.25 + 9}$$

$$= 101 \times \sqrt{25.25}$$

$$500 \text{ nV}/\sqrt{Hz}$$

For 20 kHz bandwidth, the output noise = 500 x 141
= 70,500 nV_{RMS}
= 70.5 uV_{RMS}

A 7 mV_{RMS} input signal from, say, a low impedance microphone would result in a 700 mV_{RMS} output signal and give a S/N ratio of $20 \text{ Log } \left(\frac{700}{0 \cdot 0705}\right) = 80 \text{ dB}$

Note that most dominant source in this circuit is the noise generator En. As long as the input impedance levels are kept low, then it is the noise generator En on its own that can be used as a rule of thumb for calculating absolute and comparative noise performance. For high input impedance applications a FET op amp (with virtually no input noise current) should be used.

Noise Measurements

The noise current voltage psectrums of Fig.12 were measured using a mobile analysing filter (1 Hz bandwidth) plus an RMS meter. A simpler measurement can be performed using the system shown in Fig.14. This will measure the equivalent input noise in a specified bandwidth. The low pass filter should be high order device with a steep roll off slope. If a single pole low pass filter is used, then the RMS reading should be corrected by the following equation,

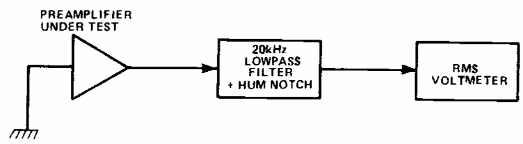


Fig.14. Noise measurement.

$$\text{'True' RMS noise (for a one pole lowpass filter)} = \sqrt{\frac{\text{measured RMS noise}}{1.57}}$$

(An RC 20 kHz low pass filter would be made from an 820 ohms resistor with a 10 n capacitor to ground).

Sometimes the signal to noise ratio is quoted in dBA. This means that the noise measurement has been modified by an A weighted curved.

A short chart of op amp performance has been drawn up in Fig.17. It is difficult to compare device performance merely from the noise voltage at one frequency in the spectrum, as the noise spectrum shapes are different from device to device. It is best to refer to the manufacturing data sheet and then to actually breadboard the devices.

DEVICE	NOISE VOLTAGE AT 1kHz nV/$\sqrt{Hz}$	UNITY GAIN BANDWIDTH MHz	SLEW RATE V/uS
NATIONAL SEMICONDUCTORS LM381A	5.0 NOT OPTIMISED	15	4.7
SIGNETICS SE5534	4 0	10	13
SIGNETICS SE5534A	3.5	10	13
741	20 to 50 NOT USUALLY SPECIFIED	1 to 1 5	0.5
RAYTHEON RC4136	10	. 3	0.5
RAYTHEON RC4558	10	3	0 5
TEXAS TL071	18 *FET INPUT VERY LOW INPUT NOISE CURRENT 0.01 PA /$\sqrt{Hz}$	3	13
FERRANTI ZN459T	4.5 (R_s = 510R)	15	--

Fig.15. Op amp performance.

Biasing

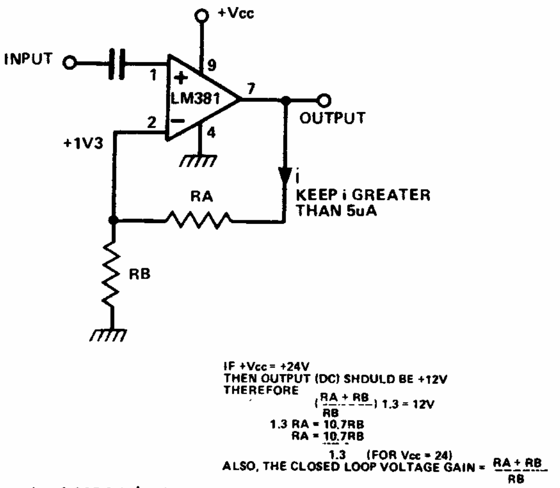


Fig.16. Biasing the LM381 (differential mode).

The base of Q1 is held at +1V3 by a pair of diodes. This preamplifier is often run from a single supply rail and a simple resistor network can be used to bias the output voltage to ½ Vcc (Fig.16). Also a single ended amplifier can be constructed (Fig.17). The resistor pair RA,RB determines the DC output level and gain. To increase the AC gain resistor RB can be shorted to ground with a series R,C network.

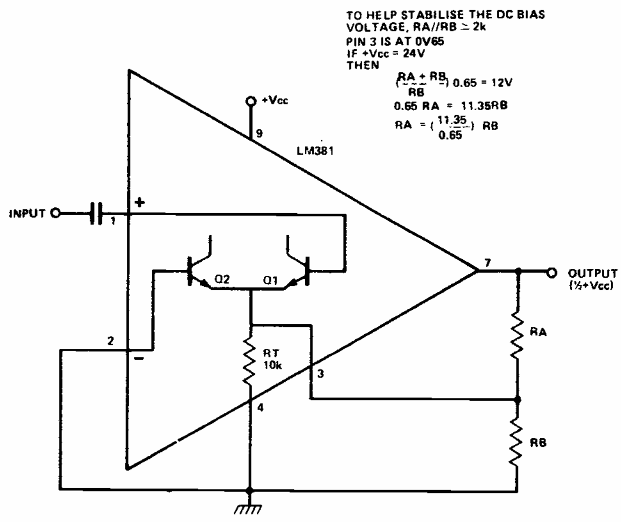


Fig.17. Single-ended biasing.

Design Example

Q. Design a preamplifier using the LM381 with a gain of 30 dB and a low frequency roll off of 20 Hz running from a +24 V power supply.

A. The design calculations are shown in Fig.18.

Record Preamplifier

When replaying a record from a magnetic cartridge it is necessary to have a preamplifier with an RIAA playback equalisation (Fig.19). A magnetic pick up generates a voltage

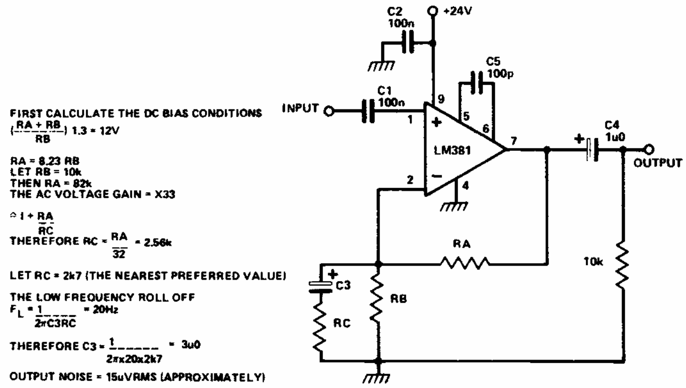


Fig.18. A 30 dB amplifier.

that is proportional to the velocity of the sideways movement of the stylus. So high frequencies produce large outputs and vice versa. Also, to assist replay electronics, the recording is given a 12 dB de-emphasis from 500 Hz to 2120 Hz. Thus, to restore a flat output it is necessary to equalise the signal from the pick up with an RIAA curve. As a rule of thumb,

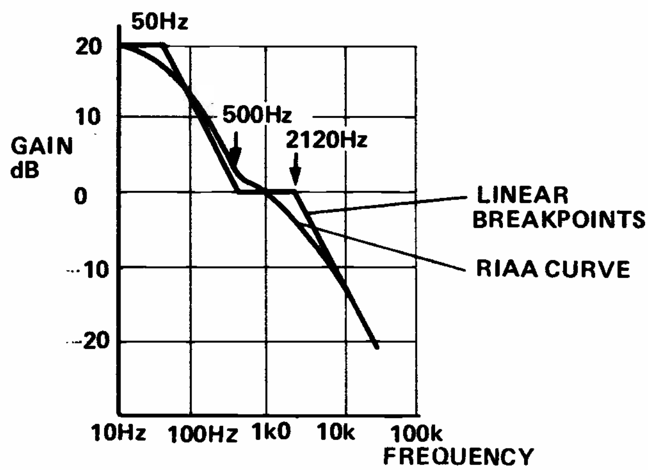


Fig.19. RIAA equalisation.

a typical magnetic pick up will generate 5 mV_{RMS} at 1 kHz, although the recording level and make of pick up will effect this figure.

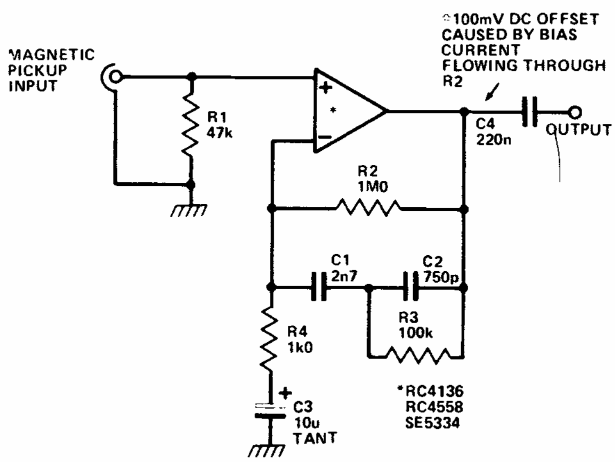


Fig.20. An RIAA-equalised preamplifier.

Low frequency gain $= \frac{R2}{R4} = 60$ dB

Low frequency rolloff $= \frac{1}{2\pi R4C3} = 15$ Hz

50 Hz breakpoint $\frac{1}{2\pi R2C1}$

The gain drops at −6 dB/octave beyond this point

500 Hz breakpoint $\frac{1}{2\pi R3C1}$

Now the gain remains constant until the next breakpoint

AC gain at 1 kHz $= \frac{R3}{R4} = 40$ dB

2120 Hz breakpoint $\frac{1}{2\pi R3C2}$

The gain now falls at −6 dB/octave beyond this frequency.

A 5 mV_{RMS} signal at 1 kHz will result in a 500 mV_{RMS} (1V4 pp) at the preamplifier output. If the amplifier is powered from ±12 V, then there is an overhead margin of

$$\left(\frac{\text{maximum output swing} = 20\text{ V}}{\text{typical swing} = 1\text{V4}}\right) = 14,$$

which is 23 dB. The noise spectrum of the op amp will be multiplied by the RIAA curve which would complicate any noise performance calculations. However, an op amp with an equivalent input noise of 0.5 uV should give a signal to noise performance of better than 76 dB which is superior than that of the disc itself. Because of the large low frequency gain care must be taken to avoid mains hum pick up. Keep the input wiring away from the mains cables and transformers, use low noise screened cable and wire this cable as close as possible to the preamplifiers.

Problems

Often a preamplifier will pick up radio signals. Usually, the radio signal is picked up by the input wiring to the preamplifier and is rectified by the transistor input stage. Then it is amplified by the rest of the audio amplifier and you end up with permanent broad band radio reception. There are several solutions that can be tried, Fig.21. A low pass filter made from an LC or an RC section will attenuate the 'pick up' interference. These devices must be physically close to the op amp. If possible use a PCB ground plane. Also, a conductive metal screen (metal foil is often used) surrounding the preamplifier will help.

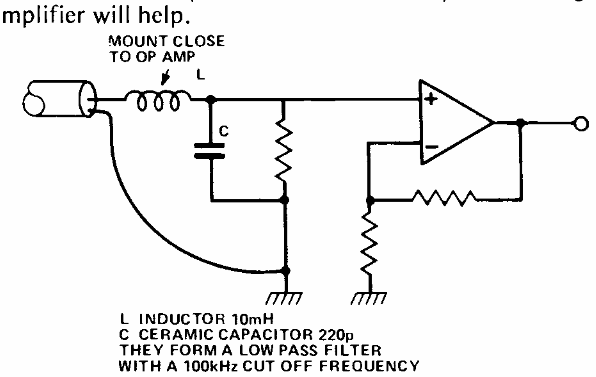
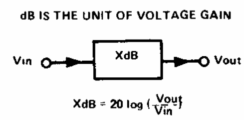


Fig.21. Removing RF interference.

Hum can also be a problem. If the hum is at 50 Hz then the source of it is probably magnetic. Check the wiring to see if any signals pass near to the mains section. Magnetic screening is difficult to implement. The best design solution is to put as much distance between the sensitive input and the mains section as possible. If it is possible try rotating the mains transformer (using a gloved hand, the other hand in your pocket). Often the size of the hum can be reduced by re-orientating the transformer. If the hum is 100 Hz, then the source is either the supply rails or the power supply layout. A larger smoothing capacitor will reduce the power supply ripple. If the hum has a sharp buzz then the problem is probably the charging current pulses in the power supply. If the layout is bad, these current pulses generate voltage pulses that get added into the ground reference voltage, thus causing the hum.

Gain

When designing an audio system it is necessary to know what the normal signal level at any point will be and also the signal gains and attenuations of various units. It has been found that the most useful way to describe levels and gains is with the logarithm decibel, Fig.22. Signal gains in dB are additive. That is, a signal passing through a series of gains of +10, +20, −30, +6 dB will end up with an overall gain of +6 dB.



dB	APPROXIMATE RATIO
+60	1000
+40	100
+20	10
+10	3
+ 6	2
+ 3	1.4
+ 0	1
- 3	0.7
− 6	0.5
−10	0.33
-20	0.1
−40	0.01
−60	0.001

dBm	Vpp	VRMS
+20	21.9V	7.75V
+ 6	4.38V	1.55V
+ 0	2.19V	0.775V
6	1.09V	0.387V
-20	0.219V	0.077V
−40	21.9mV	7.75mV
-60	2.19mV	0.77mV

dBm IS THE UNIT OF VOLTAGE LEVEL
0dBm IS 1mW OF POWER INTO A 600R LOAD

Fig.22. The dB and dBm story.

A typical audio system is shown in Fig.23. A low impedance microphone might typically give a −50 dBm output signal which will have to be given a 56 dB (x 600) gain to bring it up to line level (about +6 dBm). The line driver should

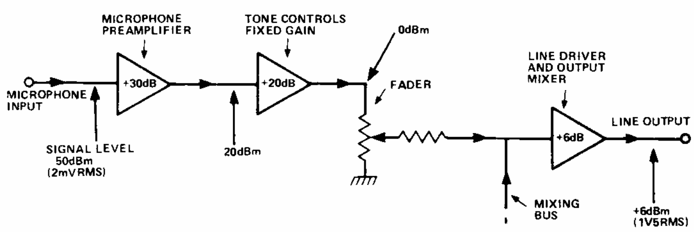


Fig.23. One channel of a mixing desk.

be capable of driving +20 dBm into 600 ohms at 20 kHz without generating significant distortion. Some line drivers are, in fact, capable of driving 30 ohm loads, but these units are small high quality power amplifiers.

2. LARGE SIGNAL, OR POWER, AMPLIFIERS

A power amplifier must deliver power into a load without generating significant distortion or bursting into oscillation or burning itself out. Power amplifiers are, however, prone to all these effects and great care is needed during their design.

Power

Power is measured in watts and is defined as the product of $V_{RMS} \times I_{RMS}$ (Fig.1c), where RMS is the equivalent DC value. That is, a 2 Vpp sinewave has the DC value of 0V7. The 2 Vpp sinewave will generate as much heat in a load as a 0V7 DC voltage. The chart in figure 1b shows the power dissipated in a load against drive voltage. If you have an amplifier that has a maximum output voltage swing of ±10 V, then the maximum power output will be 12.5 watts into a 4 ohm load (from Fig.1a). Note that the amplifier must be able to deliver a peak current of 2.5 Amps. Whilst dumping power into the load, the amplifier will be dissipating heat itself. Most monolithic devices are documented with design graphs of output power versus amplifier dissipation of power efficiency. These will enable you to determine the amplifier's maximum power dissipation. As a rule of thumb, this equals the maximum sinewave power that can be dumped into the load. A 10 watt amplifier may have to dissipate a maximum of 10 watts of heat, although this level of dissipation would not be normal in general use. Manufacturers information usually gives the thermal resistance of the junction to case. If this was say 3°C/watt then a 10 watt dissipation would raise the junction temperature by 30° above ambient (25° raising to 55°C). This is only true if the case temperature remains at ambient temperature, that is if the case is contact with an infinite heat sink. The heatsink may be anything from nothing (free air dissipation) to near infinite. It is important that the amplifier chip junction does not get very hot (above 100°C). The power chips of the amplifier age very quickly at elevated temperatures suffering from deteriorating characteristics and a short life time. This is why power amplifiers and power supplies are common failures in equipment. When the chip is heated up it expands. The chip is glued to its case and so by expanding it stresses the glue and eventually causes it to fracture. This thermal cycling increases the thermal resistance of the chip to the case and so the chip ends up operating at an even higher temperature.

Other heatsinking materials are used in the construction of power devices such as heat conducting plastics and pastes (Beryllium oxide). Manufacturers often provide design graphs of maximum dissipation versus temperature for various heatsink thermal resistances. These help to select an acceptable heatsink. Often if you are using the chassis as a heatsink it is impossible to calculate the temperature rise and it has to be done by trial and error. (As a rule of thumb I am satisfied if I can place my finger on the device for 5 seconds indicating that case temperature is no more than 80°C).

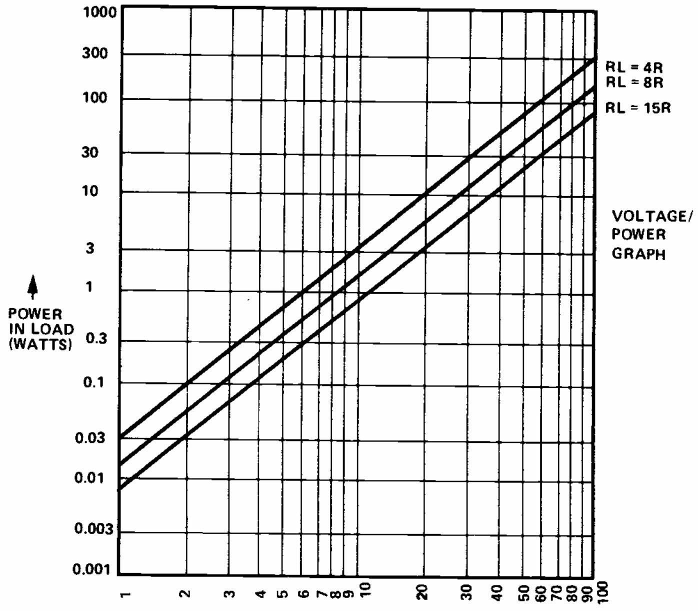
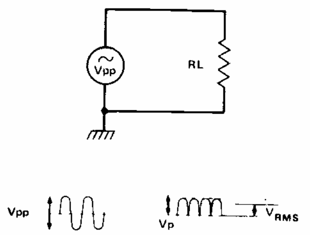


Fig.1a. (above) Graph of voltage against power.

Vpp ACROSS RL	4R	8R	15R
1V	0.031	0.015	0.008
3V	0.28	0.140	0.075
10V	3.125	1.562	0.833
30V	28.15	14.062	7.50
100V	312.5	156.25	83.33

RL — POWER IN RL

Fig.1b. (left) The power dissipated in a load against drive voltage.



$V_p = V_{pp} \times \frac{1}{2}$

$V_{RMS} = V_{pp} \times \frac{1}{2} \times - = \frac{V_{pp}}{2.828} = 0.35\,V_{pp}$

$POWER = V_{RMS} \times I_{RMS} = \frac{V_{RMS}^2}{RL}$

$POWER = \frac{V_{pp}^2}{8 \times RL}$

Fig.1c. (left) The measurement of power.

Distortion

A simple power amplifier is shown in figure 2. An op amp provides the voltage gain and a NPN, PNP transistor pair forms a current amplifier. Any distortions or nonlinearities are ironed out by the surplus loop gain of the op amp. The transfer characteristic of Q1,Q2 shows that there is a dead zone of

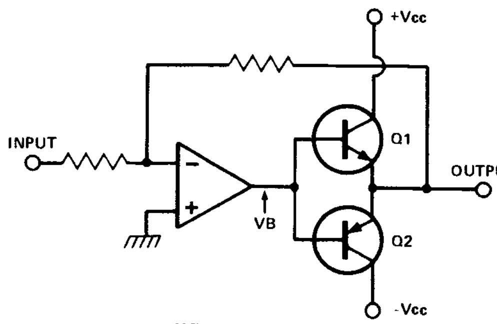


Fig.2. A simple power amplifier.

±0V6 where neither transistor is ON. If a low frequency sinewave is connected to the input (the op amp having a surplus loop gain of 1000 at this frequency) then the output will be a sinewave with a small amount of crossover distortion. The distortion level will be

$$\frac{\pm 0V6}{1000} = \pm 0.6\ mV.$$

However, as the frequency is increased, the surplus loop gain will decrease causing the distortion to rise. At higher frequencies the slew rate of the op amp becomes noticeable (Fig.3). When the input signal crosses 0 V the output of the op amp has to change from +0V6 to –0V6. If the slew rate is 0V5/uS then the time taken to travel the 1V2 distance is 2.6 uS. Thus a 2.6 uS chunk of the signal is missing out of each half cycle. The problem may be overcome by biasing the

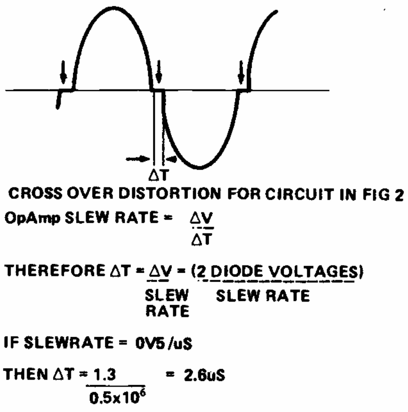


Fig.3. The effect of slew rate at higher frequencies.

two transistors so that they are both conducting. Then the crossover distortion becomes reduced to a reasonable level. The slew rate still needs to be considered. An amplifier delivering a 40 Vpp sinewave at 20 kHz has a fastest slew rate of 2V5/uS (this represents a power of 25 watts into 8 ohms). If the amplifier has not enough slew rate the output will become distorted. Manufacturers generally provide graphs of distortion (THD) versus power output and frequency (Fig. 4). For the power curve, the onset of distortion is caused by the amplifier clipping at its power rails, whereas for the frequency curve the distortion rises as the surplus loop gain falls.

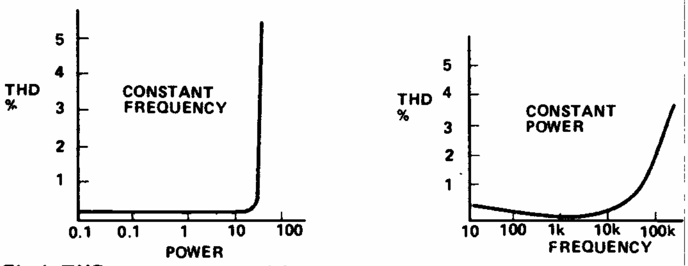


Fig.4. THD versus power and frequency.

Distortion is measured at various power levels and frequencies using the equipment shown in figure 5. A deep notch removes the test sinewave leaving behind the distortion products. Generally 0.1% THD is common for monolithic amplifiers.

Stability

Power amplifier performance is very similar to that of op amps. They have a large open loop gain that is stabilised by resistive feedback (Fig.6). Note that the amplifier described in this graph is normally inverting (180° DC phase shift) but that it suffers a phase shift as the frequency increases. The phase shift at the unity gain frequency is shown as the phase margin. If the phase margin falls to zero anywhere before the unity gain frequency then the amplifier will oscillate. This is simply because the loop phase shift will be zero at a loop gain of greater than unity, which are the conditions for an oscillation. A large phase margin is desireable.

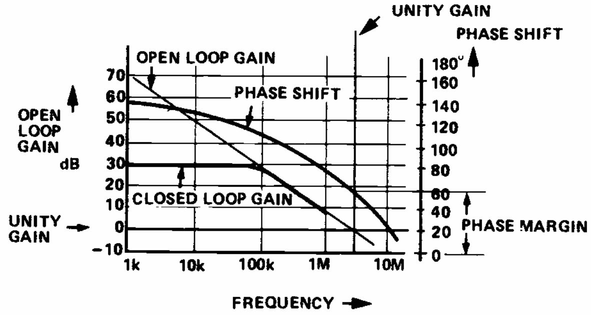


Fig.6. Gain and phase response.

Various techniques are available for preventing instability. Often a series RC network is connected from the amplifiers output to ground. This reduces the high frequency gain and thus reduces the unity gain frequency. Local power supply decoupling should be used. Current loops, whereby the output current generates a voltage in the signal ground wire which gets fed back to the input can cause bursts of high frequency oscillation. Also high input impedances can capacitively pick up the output signal and burst into oscillation. Increasing the closed loop gain may prevent some forms of instability, swearing will not! Figure 7 shows some typical instability problems.

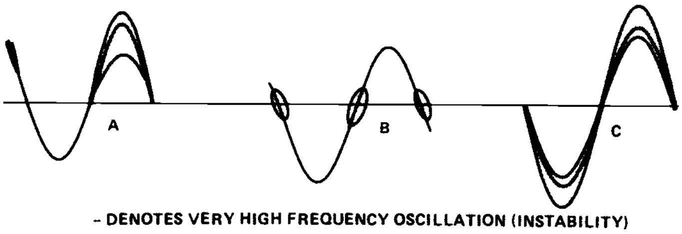


Fig.7. Instability; A, on one half cycle; B, at crossover; C, at all points.

(a) The sinewave oscillates on one half cycle. Really a power amplifier is two amplifiers, one half handling positive signals, the other, negative signals. Thus it is quite possible that the amplifier can be stable for negative signals but not positive ones.

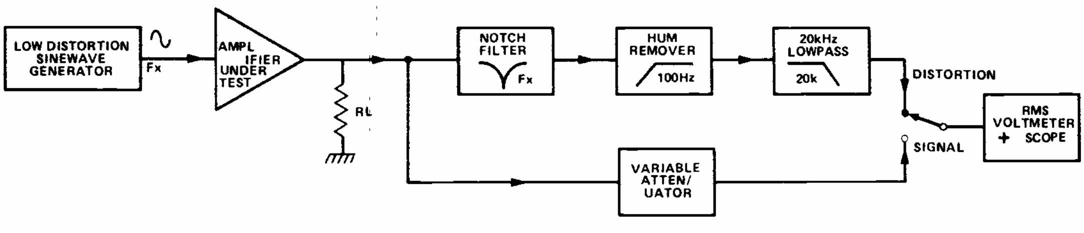


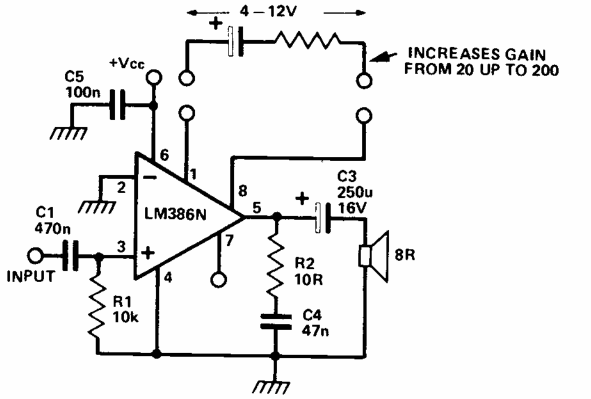
Fig.5. THD measurement.

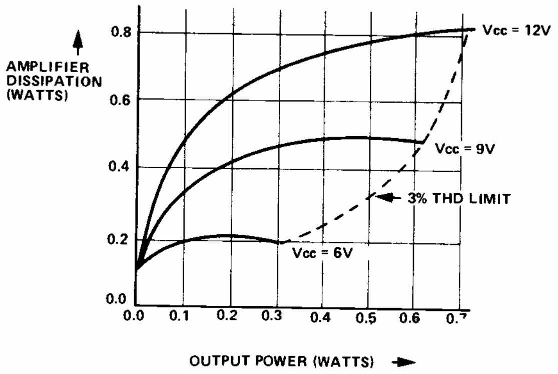
(b) The sinewave is unstable at crossover. This may be caused by the loss of feedback at crossover. Increasing the bias current may eliminate it. Alternatively it might be caused by slew limiting manifesting itself as an extra phase shift.

(c) The amplifier is never stable. Check the layout for current loops, increase the gain, increase the output CR loading, even try removing it, reduce the impedance levels.

You may not hear the effects of high frequency oscillation but it does cause RF interference and generates waste power. It can burn out output stages and even loudspeaker crossovers.

The following five design examples demonstrate how to produce a solution to an amplifier problem. It is now possible to buy an amplifier for most general purpose uses. A wide selection of monolithic devices and modules cover the 0.25 to 100 watt power range. It is rare to have the time or the ability to improve upon this range. The art of designing is to select the most suitable solution on the basis of size, cost and performance.





LM386 DEVICE DISSIPATION VERUS OUTPUT POWER INTO AN 8R LOAD

Fig.8. Low voltage, low power battery operated amplifier. The LM386N operates over a supply range of 4 to 12 V. It can deliver 0.7 watts into 8 ohms at 12 V, although, at this level, some heatsinking would be advisable. The typical battery drain is 4 mA. The voltage can be varied from 20 to 200, as shown it is 20. An AC short across pins 1 and 8 will increase the gain to 200. With a resistor in series the gain can be set to anything from 20 to 200. For gains greater than 20, a bypass capacitor (100n to ground from pin 7) should be used. Even with a supply voltage of only 4 V, there is an output voltage swing of greater than 2 V.

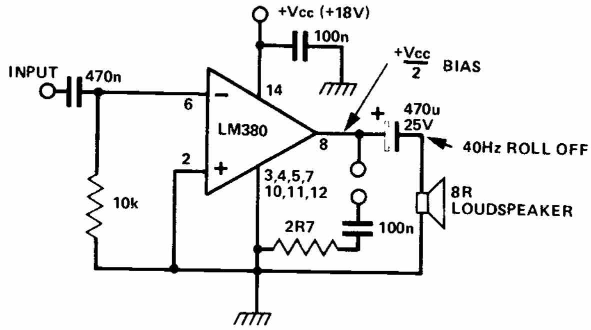
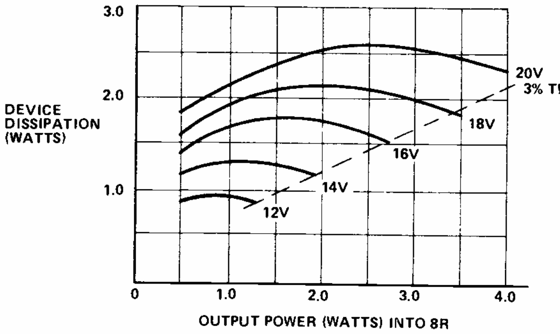
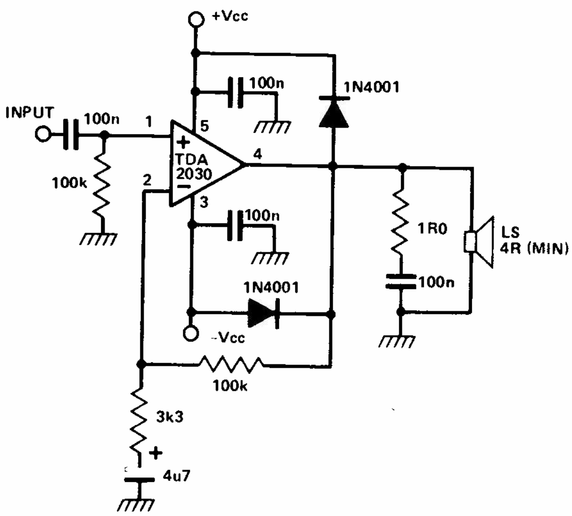


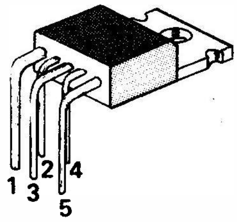
Fig.9. Medium power audio amplifier. The LM380 will work with a supply voltage range of 8 to 22 V. It can deliver 4 watts of power into 8 ohms at 20 V, although a good heatsink is needed for this level. The inputs are ground referenced and the output is automatically biased to ½ Vcc. The voltage gain is fixed at 34 dB. It also has a short circuit proof output and internal thermal limiting.





FEATURES

TO 220 PACKAGE – EASY TO HEATSINK
POWER SUPPLY RANGE ±6 TO ±18V
QUIESCENT CURRENT 40mA
POWER OUTPUT 14 WATT (4R)
9 WATT (8R)
CLOSED LOOP GAIN 30dB



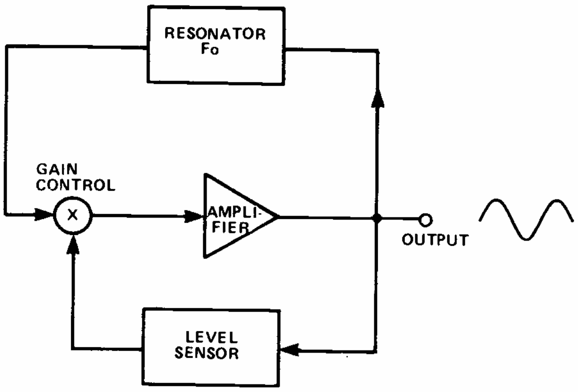
TO 220 PACKAGE

Fig.10. 9 to 14 watt amplifier. The TDA2030 is a Hi-Fi audio amplifier which has short circuit protection and thermal shutdown. It can operate from supply rails of ± 6 to ± 18 V. At a ± 14 V supply the guaranteed output power is 12 watts into 4 ohms and 8 watts into 8 ohms. Harmonic and crossover distortion is low being typically 0.05% at 1 kHz for 7 watts of power output. The recommended closed loop gain is 30 dB. The two diodes protect the amplifier from back EMF voltages from the speaker.

Experimenter's Design Notes

3. OSCILLATORS

It seems to be a fact of life that amplifiers oscillate and oscillators won't! Generally there is little difference between the two. Both devices are amplifiers with feedback. The conditions for stable sinusoidal oscillation are shown in Fig.1. The higher the Q of the resonator the more stable is the resonant frequency, and the purer the sine wave. To stabilise the signal level an automatic gain control circuit is used. This can be anything from simple diodes or thermistors to elaborate AGC systems. The smoothness with which the AGC works will determine the sine wave purity. A thermistor circuit might well introduce distortion at low frequencies by changing its resistance during one half cycle of oscillation. Very pure sinewave oscillators (better than 0.001% distortion) employ slow acting AGC systems to control the loop gain.



CONDITIONS FOR OSCILLATION

- PHASE SHIFT AROUND LOOP = 0°
- LOOP GAIN = UNITY
- IF SINEWAVE FALLS BELOW SET AMPLITUDE, GAIN INCREASES
- IF SINEWAVE INCREASES ABOVE SET AMPLITUDE, GAIN DECREASES

Fig.1. Conditions for a stable sine wave oscillator.

Wien Bridge

The well known Wien Bridge oscillator is shown in Fig.2a. A frequency sensitive feedback network is constructed from R1, C1 and R2, C2. This network has a peak in its amplitude response which also corresponds to zero phase shift. At this frequency the attenuation is x1/3 and so to ensure oscillation the amplifier must have a voltage gain of at least x3. To stabilise the amplitude a thermistor is used in the feedback loop of the op amp. As the oscillation amplitude increases, the thermistor heats up, drops in resistance and so reduces the gain. The circuit suffers from amplitude bounce when the frequency is altered. Also, the op amp phase shift, which increases with increasing frequency, must be taken into consideration when designing this oscillator. Another sine wave oscillator is shown in Fig.3. This generates both sine and cosine outputs. The circuit is a state variable filter with positive feedback (R2) to ensure oscillation and amplitude limiting (the diode bridge) to stabilise the sine wave level. The distortion may be trimmed by adjusting the amount of positive feedback. The oscillation frequency is set by RA and C.

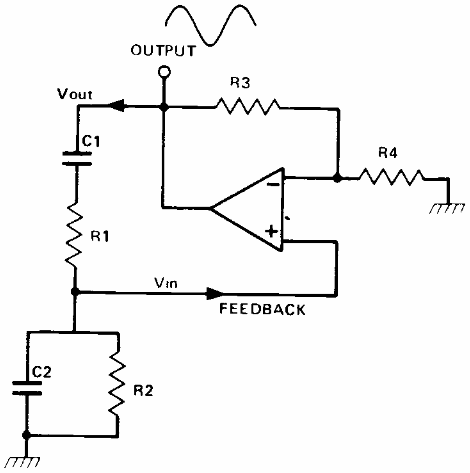


Fig.2a. A Wien Bridge oscillator.

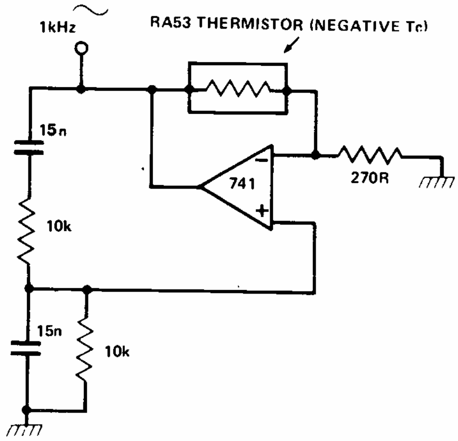


Fig.2b. To stabilise the amplifier a thermistor is used in the op amp feedback loop.

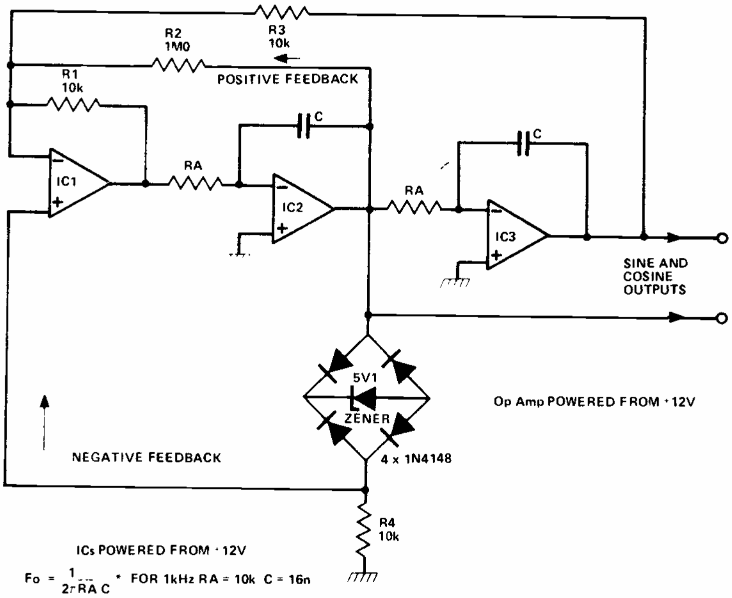


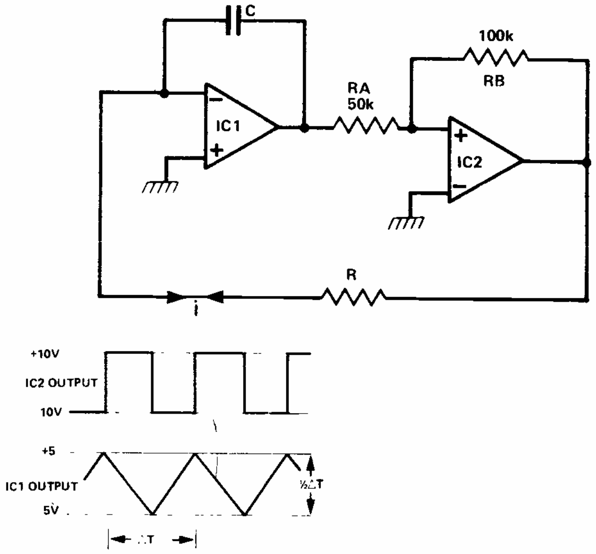
Fig.3. A state variable sine/cosine oscillator.

A triangle/square wave oscillator may be constructed from a pair of op amps (Fig.4). IC1 is an integrator, the output of which ramps up and down between the hysteresis levels set by the Schmitt trigger, IC2. If the output of IC2 can swing to ± 10 V then the hysteresis level will be $\pm \frac{10 \times RA}{RB}$ volts

The oscillation frequency and triangle symmetry is linearly proportional to the output swing of IC2. If a variable frequency oscillator is wanted, then resistor R can be connected to the wiper of a potentiometer fed from IC2 output.

Design Example

Q. Design a triangle oscillator with a 2 Vpp triangle output, oscillation frequency of 1 kHz, operating from ± 12 V power supplies, using circuit in Fig.4.



INVERTING INPUT OF IC1 IS A VIRTUAL EARTH THEREFORE $i = \frac{10V}{R}$

CURRENT CHARGES AND DISCHARGES CAPACITOR C IN THE INTEGRATOR GOVERNED BY THE RELATIONSHIP

$$C\frac{dV}{dT} = i$$

THEREFORE $\frac{dV}{dT} = \frac{i}{C} = \frac{10}{RC}$

$$\frac{dV}{dT} = \frac{\Delta V}{\Delta T}$$

SO, $\frac{1}{\Delta T} = \frac{dV}{dT} \times \frac{1}{\Delta V} = \frac{10}{RC\,\Delta V}$

WHERE $\Delta V = 20V$ AND $\frac{1}{\Delta T}$ = FREQUENCY OF OSCILLATION Fo

THEREFORE $F_o = \frac{1}{\Delta T} = \frac{10}{RC \times 20} = \frac{1}{2RC}$ Hz

Fig.4. A triangle/square wave oscillator.

A. Assume an output voltage swing at IC2 of ± 10 V.

For ± 1 V Hysteresis,

$$RA = \frac{RB}{10}$$

Let RB = 100k, then RA = 10k

$$F_o = 1\text{ kHz} = \frac{1}{\Delta T} = \frac{i}{i\Delta v} = \frac{10}{C x R x 4}$$

$$\text{Therefore } CR = \frac{10}{4x1000} = 2.5 \times 10^{-3}$$

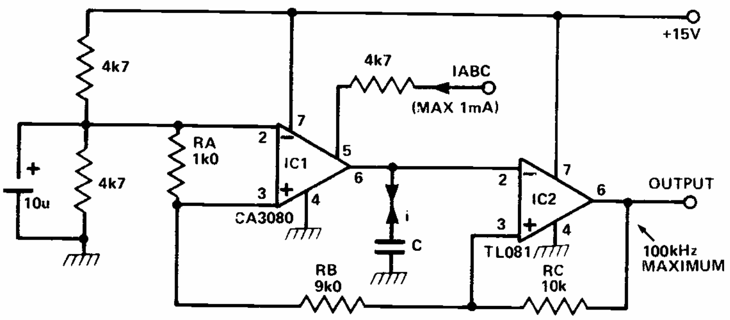
Let C = 100n

$$\text{Then } R = \frac{2.5 \times 10^{-3}}{10^{-7}} = 25k$$

Note that $\frac{\Delta V}{\Delta T}$ is 4000 V/sec. When designing this type of oscillator the slew rate of the integrator and Schmitt trigger need to be considered, although tin this case virtually any op amp will be OK.

Linear VCO

A linear voltage controlled oscillator (VCO) is shown in Fig.5. This again is a triangle square wave oscillator although the squarewave is the only buffered output. The CA3080 is used as a current source for charging and discharging C. The charging current is equal to I_{ABC}, which is true for several



CURRENT i THAT CHARGES C = ± IABC
VOLTAGE AT PIN 3 IC2 IS A 6.5Vpp SQUAREWAVE
THEREFORE, THE TRIANGLE WAVEFORM IS ALSO 6.5Vpp
EQUATION FOR CHARGING C,

$$\frac{\Delta V}{\Delta T} = \frac{i}{C} = \frac{IABC}{C}$$

$$F_o = \frac{1}{\Delta T} = \frac{IABC}{C\,\Delta V} = \frac{IABC}{C \times 6.5 \times 2}\text{ Hz}$$

* IF C = 680p, IABC = 250uA,

THEN $F_o = \frac{250 \times 10^{-6}}{680 \times 10^{-12} \times 6.5 \times 2} = 28.28$ kHz

Fig.5. A linear VCO.

decades of current. The Schmitt trigger uses a TL081 which has a slew rate of 13 V/uS. As the squarewave output voltage is 13 V then the rise and fall times are 1 uS each. This enables the VCO to run at frequencies up to 100 kHz. As this frequency is approached the VCO loses its linearity, due to time delays in the circuits.

Another VCO is shown in Fig.6. This has two buffered outputs, a triangle and a square wave. Again, the oscillation frequency is dependent on the output voltage swing of the Schmitt trigger, IC2. However, if a stabilised power supply is used this circuit behaves very well. Superior performance can be obtained by replacing Q1 with a switching FET. The abberations caused by saturation voltage and storage time are then removed. Also fast FET op amps will improve high frequency performance.

*ASSUME OUTPUT SWING OF IC2 IS ±10V
THEREFORE, IC2 SWING WILL BE ±5V

DURING HALF CYCLE A,

$i_A = (V_{in} - 1/3V_{in}$

$$i_A = \frac{(V_{in} - 1/3V_{in})}{R1 + R2}$$

$$i_A = \frac{V_{in}}{300k}$$

DURING HALFCYCLE B,

Q1 IS ON AND SO THE JUNCTION OF R1,R2 IS SHORTED TO GROUND

$$i_B = \frac{1/3\,V_{in}}{R2} = \frac{V_{in}}{300k}$$

THEREFORE, THE CURRENT THAT CHARGES AND DISCHARGES C IS THE SAME MAGNITUDE IN BOTH HALVES OF THE CYCLE

THEREFORE $\frac{\Delta V}{\Delta T} = \frac{i}{C} = \frac{V_{in}}{300k \times C}$

OSCILLATION FREQUENCY $F_o = \frac{1}{\Delta T} @ \frac{V_{in}}{300k \times C \times \Delta V} = \frac{V_{in}}{300k \times C \times 20}$

$$F_o = (\frac{V_{in}}{C} \times 0.166 \times 10^{-6})\text{ Hz}$$

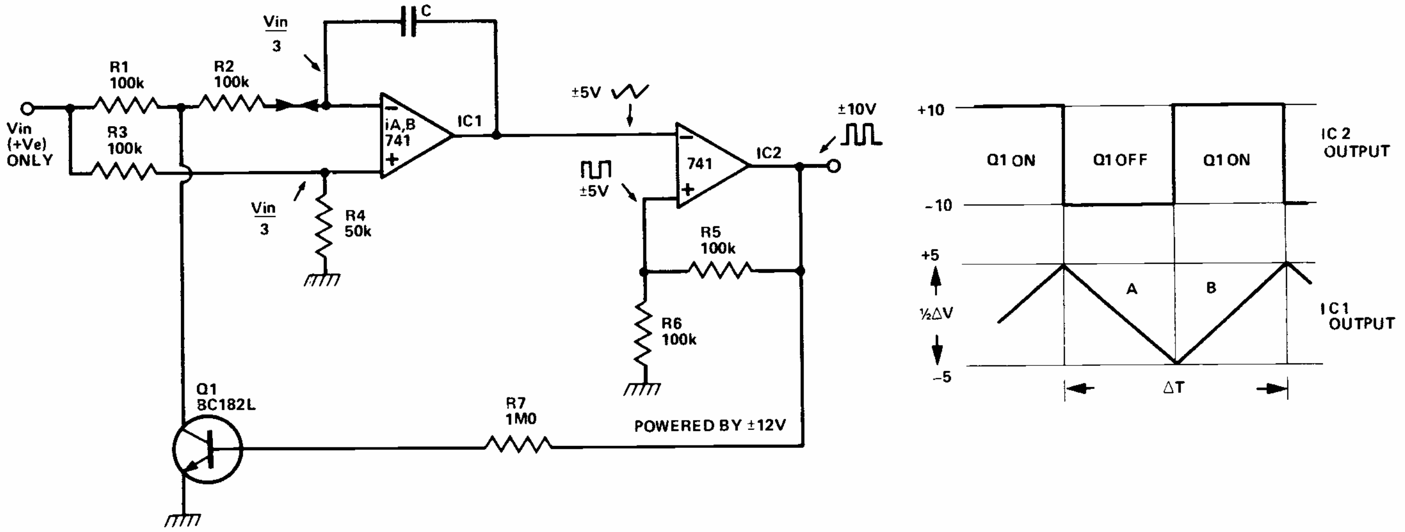


Fig.6. A linear triangle/square wave VCO.

The 555

The 555 timer chip (Fig.7) can be used as an oscillator (Fig.8). Capacitor C is charged up via RA and RB. When the voltage at pin 6,2 reaches 2/3 Vcc the discharge transistor is turned ON. When the voltage falls to 1/3 Vcc the discharge

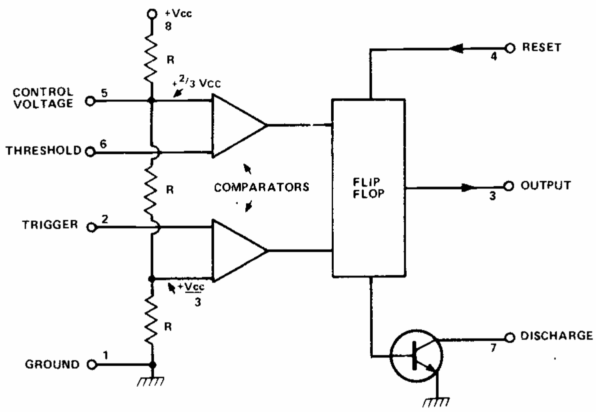


Fig.7. The timer chip.

transistor is turned OFF and the charging process repeats itself. As power supply voltage terms appear in both the numerator and demoninator of the charging equation, it drops out and so the oscillation frequency is hardly effected

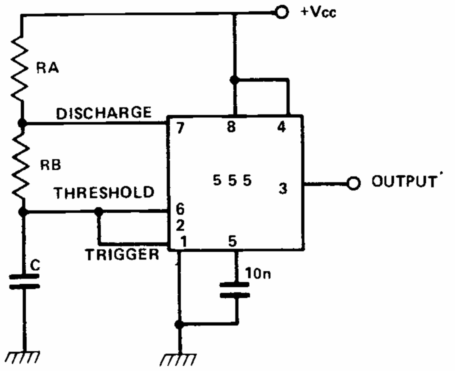
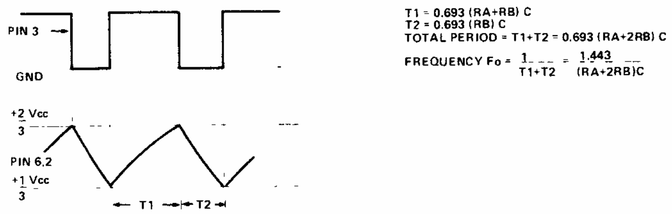


Fig.8. A simple oscillator based on the 555.

by supply voltage changes (typically 0.3%V). Also the temperature stability is good, typically 50 ppm/°C.



A low power 555 oscillator is shown in Fig.9. This employs a CMOS version of the chip which consumes a mere 120 uA. Capacitor C is slowly charged by current i and rapidly discharged by Q1.

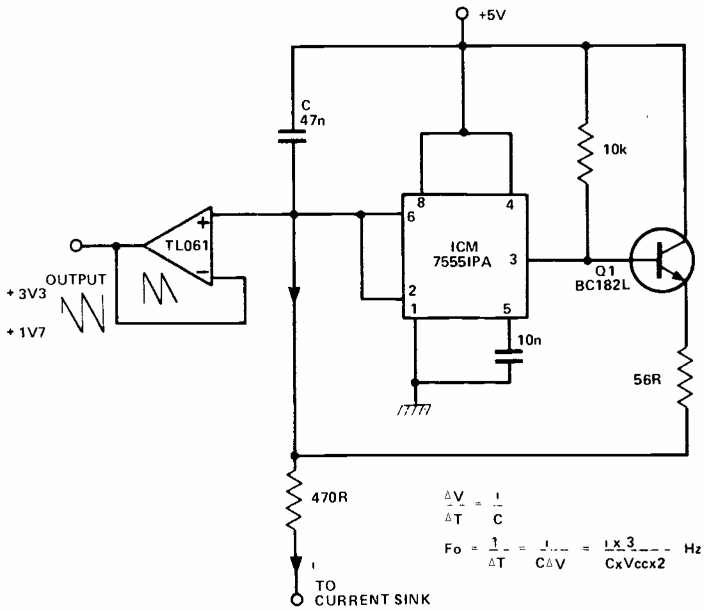


Fig.9. A linear VCO built around the ICM7555, the CMOS 555.

COMMON ABBREVIATIONS

A	Ampere or Anode
AC	Alternating Current
ACC	Automatic Chroma Control
Ae	Aerial
AF	Audio Frequency
AFC	Automatic Frequency Control
ALC	Automatic Level Control
AM	Amplitude Modulation
ANL	Automatic Noise Limiter
ATU	Aerial Tuning Unit
AVC	Automatic Volume Control
b	Base of transistor
B&S	Wire Gauge (US)
BCD	Binary Coded Decimal
C	Capacitor
c	Collector
CCD	Charge Coupled Device
CCTV	Closed Circuit Television
cgs	Centimetre-Gramme-Second
Ck	Clock
CMOS	Complementary Metal Oxide Semiconductor
CPU	Central Processing Unit
CW	Continuous Wave
D	Diode
d	Drain of FET
dB	Decibel
DC	Direct Current
DF	Direction Finding
DIL	Dual In Line
DIN	German Standards Institute
DNL	Dynamic Noise Limiter
DPDT	Double Pole Double Throw
DPST	Double Pole Single Throw
DTL	Diode Transistor Logic
DX	Long Distance
E	Voltage
ECL	Emitter Coupled Logic
EHT	Extra High Tension
EMF	Electro-Motive Force
ERP	Effective Radiated Power
F	Farad or Farenheit
f	Frequency
FET	Field Effect Transistor
FM	Frequency Modulation
G	Giga (10^9)
g	Grid or Gate
Gnd	Ground
H	Henry
HF	High Frequency
hfe	Transistor gain
HT	High Tension
Hz	Hertz
I	Current
Ib	Base Current (Transistor)
Ic	Collector current
IC	Integrated Circuit
IF	Intermediate Frequency
I^2L	Integrated Injection Logic
i/p	Input
ips	Inches per Second
K	Kilo (10^3) or Cathode
L	Inductance
LCD	Liquid Crystal Display
LDR	Light Dependent Resistor
LED	Light Emitting Diode
LF	Low Frequency
Lin	Linear
Log	Logarithmic
mA	Milliamp
mH	Millihenry
MHz	Megahertz
MOSFET	Metal Oxide Semiconductor FET
MPU	Microprocessing Unit
MSI	Medium Scale Integration
MOST	Metal Oxide Semiconductor Transistor
LS	Loudspeaker
LSI	Large Scale Integration
M	Mega (10^6)
m	Milli (10^{-3})
MPX	Multiplex
mV	Millivolt
mW	Milliwatt
n	Nano (10^{-9})
Ni-Cad	Nickel Cadmium
NR	Noise Reduction
NTSC	National Television Standards Committee
o/c	Open Circuit
o/p	Output
Op-Amp	Operational Amplifier
p	Pico (10^{-12})
PA	Power Amplifier or Public Address
PAL	Phase Alternate Line
PCB	Printed Circuit Board
pd	Potential Difference
PIL	Precision In Line
PIV	Peak Inverse Voltage
PLL	Phase Locked Loop
PROM	Programmable Read Only Memory
Ptot	Total Power Dissipation
PU	Pick Up
PUJT	Programmable Unijunction Transistor
Q	Factor of Tuned Circuit
R	Resistance
RAM	Random Access Memory
ROM	Read Only Memory
RF	Radio Frequency
RFC	Radio Frequency Choke
RMS	Root Mean Square
RTL	Resistor Transistor Logic
RX	Receiver
s	Source (FET)
s/c	Short Circuit
SCR	Silicon Controlled Rectifier
SHF	Super High Frequency
SPDT	Single Pole Double Throw
SPST	Single Pole Single Throw
SSB	Single Side Band
SSI	Small Scale Integration
SWG	Standard Wire Gauge
SWL	Short Wave Listener
SWR	Standing Wave Ratio
TRF	Tuned Radio Frequency
TTL	Transistor Transistor Logic
TVI	Television Interference
Tx	Transmitter
uF	Micro Farad
UHF	Ultra High Frequency
UJT	Unijunction Transistor
V	Volt
VA	Volt Amperes
Vcc	Supply Voltage (TTL)
VCO	Voltage Controlled Oscillator
Vdd	Supply Voltage (CMOS)
VDR	Voltage Dependent Resistor
VDU	Video Display Unit
VHF	Very High Frequency
VLF	Very Low Frequency
VMOS	Vertical Metal Oxide Semiconductor
W	Watts
X	Reactance
Xtal	Crystal
Z	Impedance

PROBLEMS?

SUFFIXES 'k', 'm', 'M' etc after component values indicate a numerical multiplier or divider – thus

Multipliers

k = X 1000
M = X 1000 000
G = X 1000 000 000
T = X 1000 000 000 000

Dividers

m = ÷ 1000
u = ÷ 1000 000
n = ÷ 1000 000 000
p = ÷ 1000 000 000 000

Where the numerical value includes a decimal point the traditional way of showing it was, for example, 4.7k. Experience showed that printing errors occurred due to accidental marks being mistaken for decimal points. The Standard now calls for the ex-suffix to be used in place of the decimal point. Thus a 4.7 k resistor is now shown as 4k7. A 2.2 uF capacitor is now shown as 2u2 etc.

Some confusion still exists with capacitor markings. Capacitors used to be marked with multiples or sub-multiples of microfarads – thus 0.001 uF, 470 uF etc. Markings are now generally in sub-multiples of a Farad. Thus –

1 microfad (1u) $=1\times10^{-6}$F
1 nanofarad (1n) = 1×10^{-12}F
1 picofarad (1p) = 1×10^{-12}F

0V on our circuits means the same as –ve (an abbreviation for 'negative').

Unless otherwise specified all components in our drawings are shown as seen from above – note however that component manufacturers often show them as seen looking *into* the pins.

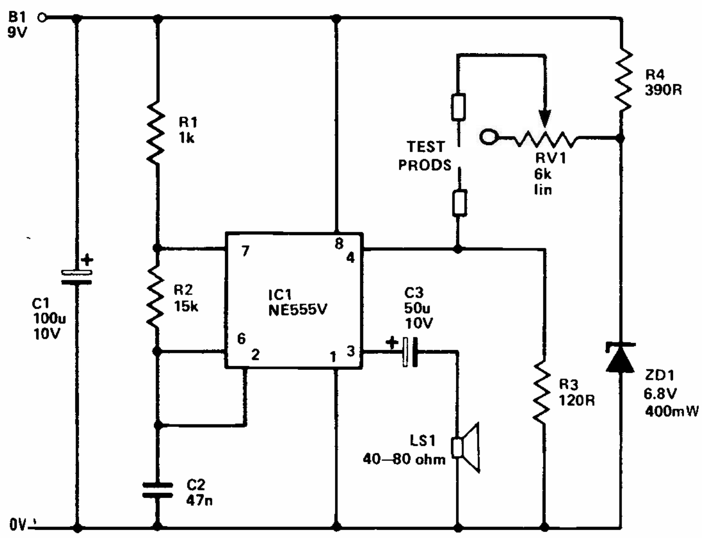
Pin numbering of ICs – with the IC held so that the pins are facing away from you and with the small cut-out downwards pins are numbered anti-clockwise starting with pin number 1 at bottom right.

The thin line on a battery schematic drawing is positive – (+ve or just +).

If a circuit won't work the most probable causes of trouble in the most probable order of occurrence are:–

(a) Components inserted the wrong way round or in the wrong places.
(b) Faulty soldering.
(c) Bridges of solder between tracks (particularly with Veroboard) – breaks in Veroboard omitted – and/or whiskers of material bridging across Veroboard breaks.
(d) Faulty components.

Continuity Tester.



A common failing of simple continuity tester circuits is that they will give an indication of continuity between the test prods when there may actually be a resistance of a few hundred ohms or more. This is often of no importance, but it can sometimes give misleading results. This simple design can be adjusted so that it will not respond to resistances of more than a few ohms.

The circuit is basically a standard 555 astable operating at a frequency of about 800Hz and feeding a high impedance speaker. However, reset terminal pin 4 is tied to the negative supply rail by R3, and this blocks the astable action. Pin 4 must be taken positive by about 0V5 or more in order to produce an audio output.

RV1 is adjusted so that with the test prods shorted together there is only just sufficient voltage at pin 4 to enable oscillation to take place. Therefore, with genuine continuity between the test prods the unit will produce an audio output, but with a resistance of more than about 7 or 8 ohms in circuit, the voltage at pin 4 will be inadequate due to the voltage drop across this resistance. RV1 is fed from a stabilised supply provided by R4 and D1 so that minor variations in the supply voltage do not necessitate readjustment of RV1. Occasional readjustment of RV1 may still be needed if it is critically adjusted for optimum discrimination.

Note that the circuit will consume power when the test produs are not connected (about 6 mA.) so on/off switch SW1 is required.

Transistor Checker

This device enables transistors to be quickly tested and shows whether or not they are serviceable.

If we consider the circuit in the NPN mode first, it is basically a standard astable multivibrator with the test transistor as one of the transistors in the circuit. The test device is biased by R1 and has R3 as its collector load resistance. R2 shunts the base emitter terminals of the test device, but is too high in value to have any significant effect on the circuit. Q1 is used in the other section of the multivibrator, and it is biased by R4. LED indicator D1 and current limiting resistor R5 act as its collector load. C2 and C3 provide the cross coupling between the two stages.

If the test device is functional, the circuit will of course oscillate and the specified values give an operating frequency of about 2 to 3 Hertz. As Q1 switches on and off, D1 will be seen to flash on and off. If the test device is faulty, R4 will simply bias Q1 into conduction, resulting in D1 remaining switched on while SW2 is depressed and power is supplied to the circuit.

With SW1 switched to the "PNP" mode, the supply to the test device is altered to the correct polarity. R2 now biases the test device; R1 has no significant effect on the circuit. Although a multivibrator would not normally use one PNP and one NPN device, the circuit operates very much as before.

The current consumption of the unit is about 6 mA. Three small wander sockets mounted in a triangular arrangement can be used as the test sockets and provided they are mounted very close together, most small transistors will plug in without difficulty. A set of three test leads terminated in crocodile clips can be used to connect to those that will not.