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R. M. MARSTON

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PREFACE

Integrated circuits are the most important semiconductor devices in use today. They are compact, relatively easy to use, and are less expensive than their discrete component equivalents. This fully revised second edition of this volume contains an entirely practical introduction to five specific and popular types of linear integrated circuits, and shows one hundred and ten constructional projects in which they can be used. These projects range from simple low-level amplifiers to precision measuring and indicating circuits. The volume will, like other books in the '110 Projects' series, be of equal interest to the amateur, the student, and the professional engineer.

The text is divided into five chapters, each of which describes a number of projects that can be built around a single i.c. type; all i.c.s are popular types, and are readily available in all parts of the western world. Care has been taken throughout the volume to ensure that all circuit diagrams convey constructional as well as theorectical information: Where appropriate, the diagrams are drawn around the i.c. outline to convey suggested layout information.

All projects described in the volume have been designed, built, and fully evaluated by the author. Two chapters (on the 741 op-amp and the 555 timer) have appeared in the magazine *Electronics Today International,* and one chapter (on the XR-2206 function generator) has appeared in the American magazine *Radio-Electronics*, so most of the projects have been further evaluated by many thousands of readers.

The outlines and pin connections of all semiconductors mentioned in the volume are given in the Appendix, as an aid to construction. Unless otherwise stated, all resistors used in the circuits are standard half-watt types.

In case of difficulty, UK readers can obtain (at the time of writing) all semiconductor devices mentioned in this volume from Arrow Electronics Ltd, Leader House, Coptfold Road, Brentwood, Essex.

CHAPTER 1

TYPE 741 OP-AMP PROJECTS

Operational amplifiers (op-amps) can be simply described as high-gain direct-coupled voltage amplifier 'blocks' that have a single output terminal but have both inverting and non-inverting input terminals. Op-amps can readily be used as inverting, non-inverting, and differential amplifiers in both a.c. and d.c. applications, and can easily be made to act as oscillators, tone filters, and level switches, etc.

Op-amps are readily available in integrated circuit form, and as such act as one of the most versatile building blocks available in electronics today. One of the most popular i.c. op-amps presently available is the device that is universally known as the type 741 op-amp.

This present chapter describes the basic features of this device and shows a wide variety of practical circuits in which it can be used.

Basic op-amp characteristics and circuits

In its simplest form, an op-amp consists of a differential amplifier followed by offset compensation and output stages, as shown in Fig. 1.1a. The differential amplifier has inverting and non-inverting input terminals, has a high-impedance (constant current) tail to give a high input impedance and a high degree of common mode signal rejection, and has a high-impedance (constant current) load to give a high degree of signal voltage stage gain. The output of the differential amplifier is fed to a direct-coupled offset compensation stage, which effectively reduces the output offset voltage of the differential amplifier to zero volts under quiescent conditions, and the output of the compensation stage is fed to a simple complementary emitter follower output stage, which gives a low output impedance.

Op-amps are normally powered from split power supplies, providing +ve, –ve, and common (zero volt) supply rails, so that the output of the op-amp can swing either side of the zero volts value, and can be set

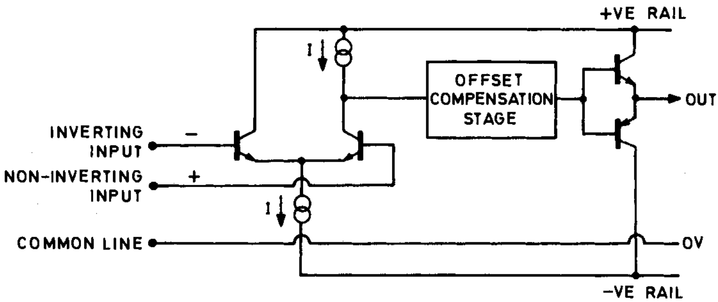


Fig. 1.1a. Simplified op-amp equivalent circuit

at a true zero volts when zero differential voltage is applied to the circuit's input terminals. The op-amp input terminals can be used independently (with the unused terminal grounded) or simultaneously, enabling the device to function as an inverting, non-inverting, or differential amplifier. Since the op-amp is direct coupled throughout, it

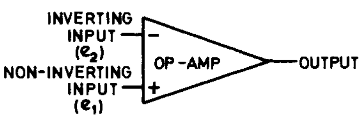


Fig. 1.1b. Basic op-amp symbol

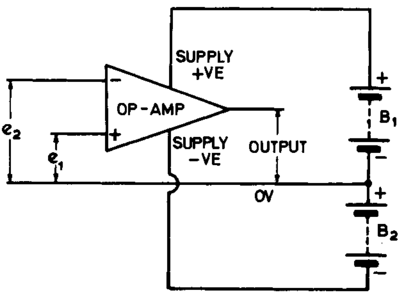


Fig. 1.1c. Basic supply connections of an op-amp

can be used to amplify both a.c. and d.c. input signals. Typically, op-amps give basic low-frequency voltage gains of about 100 000 between input and output, and have input impedances of 1MΩ or greater at each input terminal.

Fig. 1.1b shows the symbol that is commonly used to represent the op-amp, and Fig. 1.1c shows the basic supply connections that are used with the device. Note that both input and output signals of the op-amp are referenced to the ground or zero volt line.

The output signal voltage of the op-amp is proportional to the *differential* signal between its two input terminals, and is given by

$$e_{out} = A_o\ (e_1 - e_2)$$

where A_o = the open-loop voltage gain of the op-amp (typically 100 000), e_1 = signal voltage at the non-inverting input terminal, and e_2 = signal voltage at the inverting input terminal.

Thus, if identical signals are simultaneously applied to both input terminals, the circuit will (ideally) give zero signal output. If a signal is applied to the inverting terminal only, the circuit gives an amplified and inverted output. If a signal is applied to the non-inverting terminal only, the circuit gives an amplified but non-inverted output. By using external negative feedback components, the stage gain of the op-amp circuit can be very precisely controlled.

Fig. 1.2a shows a very simple application of the op-amp. This particular circuit is known as a differential voltage comparator, and has

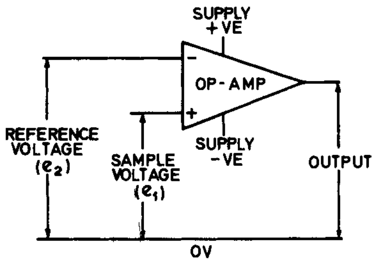


Fig. 1.2a. Simple differential voltage comparator circuit

a fixed reference voltage applied to the inverting input terminal, and a variable test or sample voltage applied to the non-inverting terminal. When the sample voltage is more than a few hundred microvolts above the reference voltage the op-amp output is driven to saturation in a positive direction, and when the sample is more than a few hundred microvolts below the reference voltage the output is driven to saturation in the negative direction.

Fig. 1.2b shows the voltage transfer characteristics of the above circuit. Note that it is the magnitude of the differential input voltage that dictates the magnitude of the output voltage, and that the absolute values of input voltage are of little importance. Thus, if a 1V reference is used and a differential voltage of only 200μV is needed to switch the output from a negative to a positive saturation level, this change can be caused by a shift of only 0.02% on a 1V signal applied to the sample input. The circuit thus functions as a precision voltage comparator or balance detector.

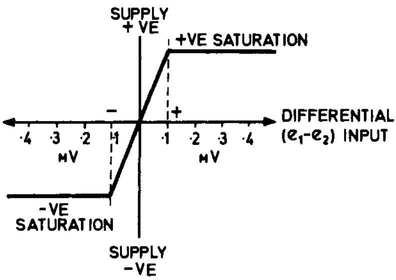


Fig. 1.2b. Transfer characteristics of the differential voltage comparator circuit

The op-amp can be made to function as a low-level inverting d.c. amplifier by simply grounding the non-inverting terminal and feeding the input signal to the inverting terminal, as shown in Fig. 1.3a. The op-amp is used 'open-loop' (without feedback) in this configuration,

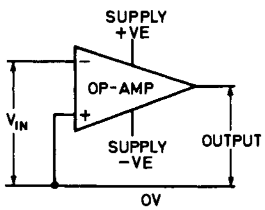


Fig. 1.3a. Simple open-loop inverting d.c. amplifier

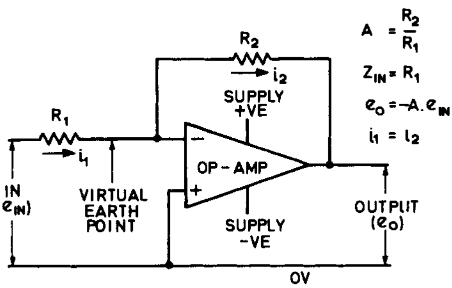


Fig. 1.3b. Basic closed-loop inverting d.c. amplifier

and thus gives a voltage gain of about 100 000 and has an input impedance of about 1MΩ. The disadvantage of this circuit is that its parameters are dictated by the actual op-amp, and are subject to considerable variation between individual devices.

A far more useful way of employing the op-amp is to use it in the closed-loop mode, i.e., with negative feedback. Fig. 1.3b shows the method of applying negative feedback to make a fixed-gain inverting d.c. amplifier. Here, the parameters of the circuit are controlled by feedback resistors R_1 and R_2. The gain, A, of the circuit is dictated by the ratios of R_1 and R_2, and equals R_2/R_1. The gain is virtually independent of the op-amp characteristics, provided that the open-loop gain (A_0) is large relative to the closed-loop gain (A). The input impedance of the circuit is equal to R_1, and again is virtually independent of the op-amp characteristics.

It should be noted at this point that although R_1 and R_2 control the gain of the complete circuit, they have no effect on the parameters of the actual op-amp, and the full open-loop gain of the op-amp is still available between its inverting input terminal and the output. Similarly, the inverting terminal continues to have a very high input impedance, and negligible signal current flows into the inverting terminal. Consequently, virtually all the R_1 signal current also flows in R_2, and signal currents i_1 and i_2 can be regarded as being equal, as indicated in the diagram.

Since the signal voltage appearing at the output terminal end of R_2 is A times greater than that appearing at the inverting terminal end, the current flowing in R_2 is A times greater than that caused by the inverting terminal signal only. Consequently, R_2 has an apparent value of R_2/A when looked at from its inverting terminal end, and the R_1–R_2 junction thus appears as a low-impedance *virtual earth* point.

It can be seen from the above description that the Fig. 1.3b circuit is very versatile. Its gain and input impedance can be very precisely controlled by suitable choice of R_1 and R_2, and are unaffected by

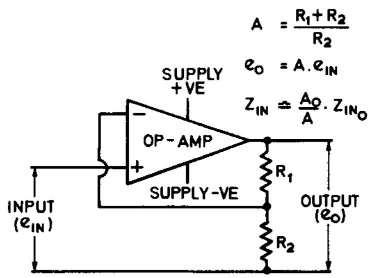


Fig. 1.4a. Basic non-inverting d.c. amplifier

variations in the op-amp characteristics. This is also true of the non-inverting d.c. amplifier circuit shown in Fig. 1.4a. In this case the voltage gain is equal to $(R_1 + R_2)/R_2$, and the input impedance is approximately equal to $(A_0/A)Z_{in_0}$, where Z_{in_0} is the open-loop input impedance of the op-amp. A great advantage of this circuit is that it has a very high input impedance.

The op-amp can be made to function as a precision voltage follower by connecting it as a unity-gain non-inverting d.c. amplifier, as shown in Fig. 1.4b. In this case the input and output voltages of the circuit are identical, but the input impedance is very high and is roughly equal to $A_O \times Z_{in_O}$.

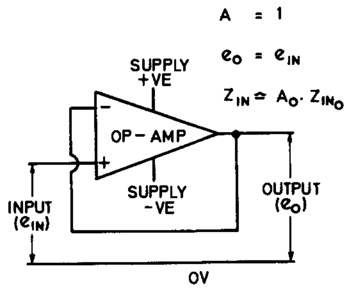


Fig. 1.4b. Basic unity gain d.c. voltage follower

The basic op-amp circuits of Figs. 1.2a to 1.4b are shown as d.c. amplifiers, but can readily be adapted for a.c. use. Op-amps also have many applications other than as simple amplifiers. They can easily be made to function as precision phase splitters, as adders or subtractors, as active filters or selective amplifiers, as precision half-wave or full-wave rectifiers, and as oscillators or multivibrators, etc. A whole range of useful applications are described later in this chapter.

Op-amp parameters

An ideal op-amp would have an infinite input impedance, zero output impedance, infinite gain and infinite bandwidth, and would give perfect tracking between input and output. Practical op-amps fall far short of this ideal, and have finite gain, bandwidth, etc., and give tracking errors between the input and output signals. Consequently, various performance parameters are detailed on op-amp data sheets, and indicate the measure of 'goodness' of the particular device in question. The most important of these parameters are detailed below.

Open-loop voltage gain, A_O. This is the low-frequency voltage gain occuring directly between the input and output terminals of the op-amp, and may be expressed in direct terms or in terms of dB. Typically, d.c. gain figures of modern op-amps are 100 000, or 100dB.

Input impedance, Z_{in}. This is the impedance looking directly into the input terminals of the op-amp when it is used open-loop, and is usually expressed in terms of resistance only. Values of 1MΩ are typical of modern op-amps with bipolar input stages, while f.e.t. input types have impedances of a million MΩ or greater.

Output impedance, Z_o. This is the output impedance of the basic op-amp when it is used open-loop, and is usually expressed in terms of resistance only. Values of a few hundred ohms are typical of modern op-amps.

Input bias current, I_b. Many op-amps use bipolar transistor input stages, and draw a small bias current from the input terminals. The magnitude of this current is denoted by I_b, and is typically only a fraction of a microamp.

Supply voltage range, V_s. Op-amps are usually operated from two sets of supply rails, and these supplies must be within maximum and minimum limits. If the supply voltages are too high the op-amp may be damaged, and if the supply voltages are too low the op-amp will not function correctly. Typical supply limits are ±3V to ±15V.

Input voltage range, $V_{i(max)}$. The input voltage to the op-amp must never be allowed to exceed the supply line voltages, or the op-amp may be damaged. $V_{i(max)}$ is usually specified as being one or two volts less than V_s.

Output voltage range, $V_{o(max)}$. If the op-amp is over driven its output will saturate and be limited by the available supply voltages, so $V_{o(max)}$ is usually specified as being one or two volts less than V_s.

Differential input offset voltage, V_{io}. In the ideal op-amp perfect tracking would exist between the input and output terminals of the device, and the output would register zero when both inputs were grounded. Actual op-amps are not perfect devices, however, and in practice slight imbalances exist within their input circuitry and effectively cause a small offset or bias potential to be applied to the input terminals of the op-amp. Typically, this *differential input offset voltage* has a value of only a few millivolts, but when this voltage is amplified by the gain of the circuit in which the op-amp is used it may be sufficient to drive the op-amp output to saturation. Because of this, most op-amps have some facility for externally nulling out the offset voltage.

Common mode rejection ratio, c.m.r.r. The ideal op-amp produces an output that is proportional to the difference between the two signals applied to its input terminals, and produces zero output when identical signals are applied to both inputs simultaneously, i.e., in common mode. In practical op-amps, common mode signals do not entirely cancel out, and produce a small signal at the op-amp output terminal. The ability of the op-amp to reject common mode signals is usually expressed in terms of common mode rejection ratio, which is the ratio of op-amp gain with differential signals to op-amp gain with common mode signals. C.m.r.r. values of 90dB are typical of modern op-amps.

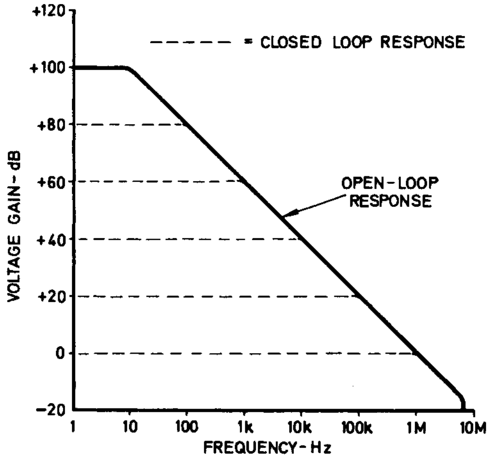


Fig. 1.5. Typical frequency response curve of the 741 op-amp

Transition frequency, f_T. An op-amp typically gives a low-frequency voltage gain of about 100dB, and in the interest of stability its open-loop frequency response is tailored so that the gain falls off as the frequency rises, and falls to unity at a transition frequency denoted f_T. Usually, the response falls off at a rate of 6dB per octave or 20dB per decade. Fig. 1.5 shows the typical response curve of the type 741 op-amp, which has an f_T of 1MHz and a low frequency gain of 100dB.

Note that, when the op-amp is used in a closed-loop amplifier circuit, the bandwidth of the circuit depends on the closed-loop gain. If the amplifier is used to give a gain of 60dB its bandwidth is only 1kHz, and if it is used to give a gain of 20dB its bandwidth is 100kHz. The f_T figure can thus be used to represent a gain-bandwidth product.

Slew rate, S. In addition to the normal bandwidth limitations, op-amps are also subject to a phenomenon known as slew rate limiting, which has the effect of limiting the maximum rate of change of voltage at the output of the device. Slew rate is normally specified in terms of volts per microsecond, and values in the range 1V/μs to 10V/μs are common with most popular types of op-amp. One effect of slew rate limiting is to make a greater bandwidth available to small output signals than is available to large output signals.

The type 741 op-amp

Early types of i.c. op-amp, such as the well known 709 type, suffered from a number of design weaknesses. In particular, they were prone to a phenomenon known as *input latch-up*, in which the input circuitry tended to switch into a locked state if special precautions were not taken when connecting the input signals to the input terminals, and tended to self-destruct if a short circuit were inadvertently placed across the op-amp output terminals. In addition, the op-amps were prone to bursting into unwanted oscillations when used in the linear amplifier mode, and required the use of external frequency compensation components for stability control.

Table 1.1 Typical characteristics of the 741 op-amp

	Parameter	*741 value*
A_o	Open-loop voltage gain	100dB
Z_{in}	Input impedance	1MΩ
Z_o	Output impedance	150Ω
I_b	Input bias current	200mA
$V_{s,\,max}$	Maximum supply voltage	± 18V
$V_{i,\,max}$	Maximum input voltage	± 13V
$V_{o,\,max}$	Maximum output voltage	± 14V
V_{io}	Differential input offset voltage	2mV
c.m.m.r.	Common mode rejection ratio	90dB
f_T	Transition frequency	1MHz
S	Slew rate	1V/μs

These weaknesses have been eliminated in the type 741 op-amp. This device is immune to input latch-up problems, has built-in output short circuit protection, and does not require the use of external frequency compensation components. The typical performance characteristics of the device are listed in Table 1.1.

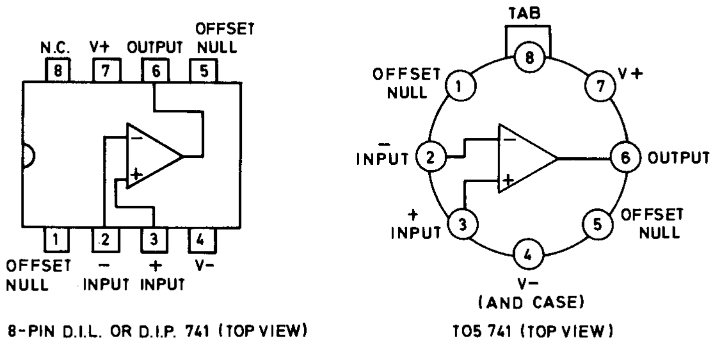


Fig. 1.6. Outlines and pin connections of the two most popular 741 packages

The type 741 op-amp is marketed by most i.c. manufacturers, and is very readily available. Fig. 1.6 shows the two most commonly used forms of packaging of the device. Throughout this chapter, all practical circuits are based on the standard 8-pin dual-in-line (DIL or DIP) version of the 741 op-amp.

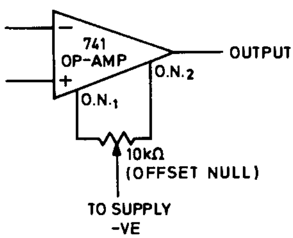


Fig. 1.7. Method of applying offset nulling to the 741 op-amp

The 741 op-amp can be provided with external offset nulling by wiring a 10kΩ pot between its two null terminals and taking the pot slider to the negative supply rail, as shown in Fig. 1.7.

Basic linear amplifier projects

Fig. 1.8 to 1.11 show a variety of ways of using the 741 op-amp in basic linear amplifier applications.

The 741 op-amp can be made to function as an inverting amplifier by grounding the non-inverting input terminal and feeding the input signal to the inverting terminal. The voltage gain of the circuit can be precisely controlled by selecting suitable values of external feedback

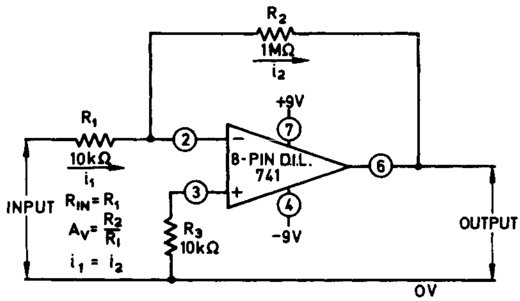


Fig. 1.8a. x100 inverting d.c. amplifier

resistance. Fig. 1.8a shows the practical connections of an inverting d.c. amplifier with a preset gain of x100: The voltage gain is determined by the ratios of R_1 and R_2, as shown in the diagram. The gain can be readily altered by using alternative R_1 and/or R_2 values. If required,

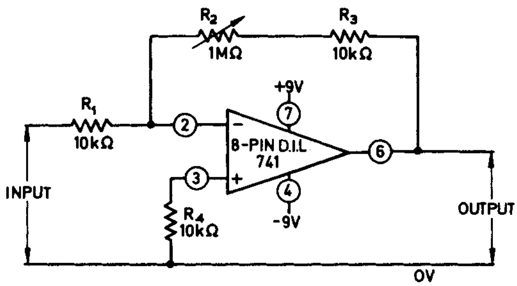


Fig. 1.8b. Variable gain (x1 to x100) inverting d.c. amplifier

the gain can be made variable by using a series combination of a fixed and a variable resistor in place of R_2, as shown in the circuit of Fig.1.8b, in which the gain can be varied over the range x1 to x100 via R_2.

A variation of the basic inverting d.c. amplifier is shown in Fig. 1.9a. Here, the feedback connection to R_2 is taken from the output of the R_3–R_4 output potential divider, rather than directly from the output

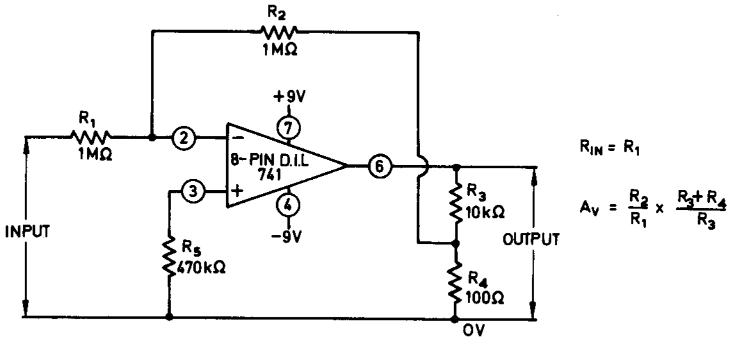


Fig. 1.9a. High impedance x100 inverting d.c. amplifier

of the op-amp, and the voltage gain is determined by the ratios of this divider as well as by the values of R_1 and R_2. The important feature of this circuit is that it enables R_1, which determines the input impedance of the circuit, to be given a high value if required, while at the same time enabling high voltage gain to be achieved.

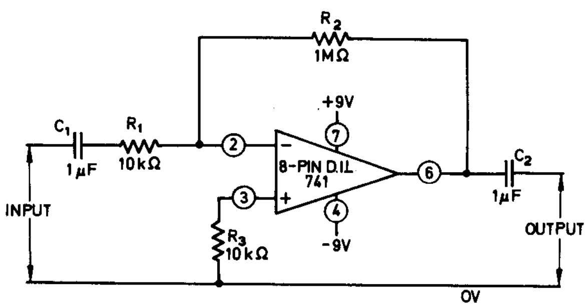


Fig. 1.9b. x100 inverting a.c. amplifier

The basic inverting d.c. amplifier can be adapted for a.c. use by simply wiring blocking capacitors in series with its input and output terminals, as shown in the x100 inverting a.c. amplifier circuit of Fig. 1.9b.

The 741 op-amp can be made to function as a non-inverting amplifier by feeding the input signal to its non-inverting terminal and applying negative feedback to the inverting terminal via a resistive potential divider that is connected across the op-amp output. Fig. 1.10a shows the connections for making a fixed gain (x100) d.c. amplifier.

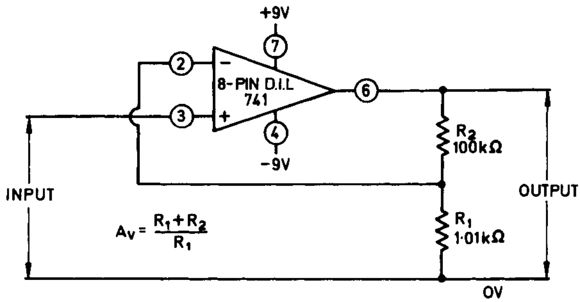


Fig. 1.10a. Non-inverting x100 d.c. amplifier

The voltage gain of the Fig. 1.10a circuit is determined by the ratios of R_1 and R_2. If R_2 is given a value of zero the gain falls to unity, and if R_1 is given a value of zero the gain rises towards infinity (but in practice is limited to the open-loop gain of the op-amp). If required, the gain can be made variable by replacing R_2 with a potentiometer and

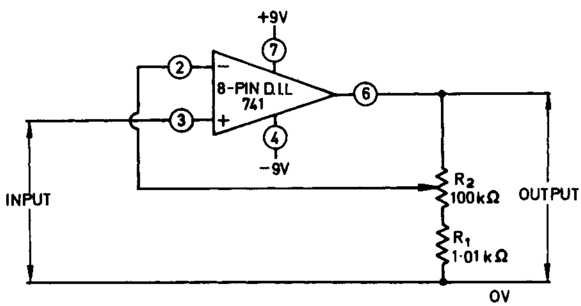


Fig. 1.10b. Non-inverting variable gain (x1 to x100) d.c. amplifier

connecting the pot slider to the inverting terminal of the op-amp, as shown in the circuit of Fig. 1.10b. The gain of this circuit can be varied over the range x1 to x100 via R_2.

A major advantage of the non-inverting d.c. amplifier is that it has a very high input resistance. In theory, the input resistance is equal to the open-loop input resistance (typically 1MΩ) multiplied by the open-loop voltage gain (typically 100 000) divided by the actual circuit voltage gain. In practice, input resistance values of hundred of megohms can readily be obtained.

The basic non-inverting d.c. circuits of Fig. 1.10 can be modified to operate as a.c. amplifiers in a variety of ways. The most obvious approach

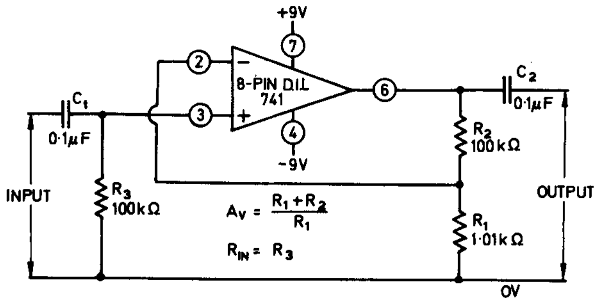


Fig. 1.11a. Non-inverting x100 a.c. amplifier

here is to simply wire blocking capacitors in series with the inputs and outputs, but in such cases the input terminal must be d.c. grounded via a suitable resistor, as shown by R_3 in the non-inverting x100 a.c. amplifier of Fig. 1.11a. If this resistor is not used the op-amp will have

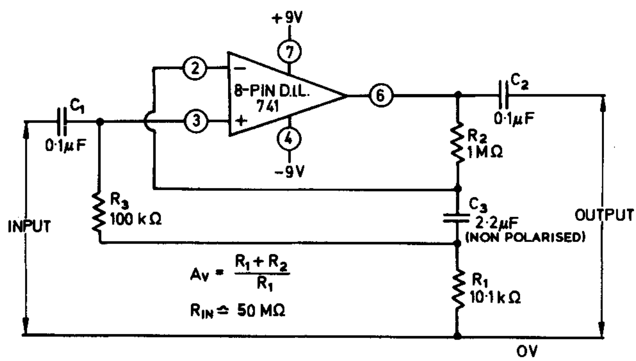


Fig. 1.11b. Non-inverting, high input impedance x100 a.c. amplifier

no d.c. stability, and its output will rapidly drift into saturation. Clearly, the input resistance of the Fig. 1.11a circuit is equal to R_3, and R_3 must have a relatively low value in the interest of d.c. stability. This circuit thus loses the non-inverting amplifier's basic advantage of high input resistance.

A useful development of the Fig. 1.11a circuit is shown in Fig. 1.11b. Here, the values of R_1 and R_2 are increased and a blocking capacitor is interposed between them. At practical operating frequencies this capacitor has a negligible impedance, so the voltage gain is still determined by the ratios of the two resistors. Because of the inclusion of the blocking capacitor, however, the inverting terminal of the op-amp is subjected to virtually 100% d.c. negative feedback from the output terminal of the op-amp, and the circuit thus has excellent d.c. stability. The low end of R_3 is connected to the C_3–R_1 junction, rather than directly to the ground line, and the signal voltage appearing at this point is virtually identical with that appearing at the non-inverting terminal of the op-amp. Consequently, identical signal voltages appear at both ends of R_3, and the apparent impedance of this resistor is increased close to infinity by bootstrap action. This circuit thus has good d.c. stability and a very high input impedance. In practice, this circuit gives a typical input impedance of about 50MΩ.

Voltage follower projects

The 741 op-amp can be made to function as a precision voltage follower by connecting it as a unity-gain non-inverting amplifier. Fig. 1.12a shows the practical connections for making a d.c. voltage follower. Here,

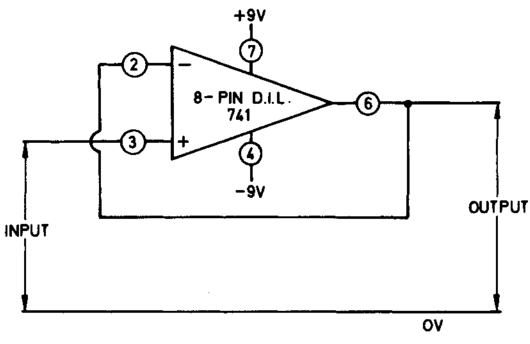


Fig. 1.12a. D.C. voltage follower

the input signal is applied directly to the non-inverting terminal of the op-amp, and the inverting terminal is connected directly to the output, so the circuit has 100% d.c. negative feedback and acts as a unity-gain non-inverting d.c. amplifier. The output signal voltage of the circuit is virtually identical to that of the input, so the output is said to 'follow' the input voltage. The great advantage of this circuit is that it has a very high input impedance (as high as hundred of megohms) and a very low output impedance (as low as a few ohms). The circuit acts effectively as an impedance transformer.

In practice the output of the basic Fig. 1.12a circuit will follow the input to within a couple of millivolts up to magnitudes within a volt or so of the supply line potentials. If required, the circuit can be made to follow to within a few microvolts by adding the offset null facility to the op-amp.

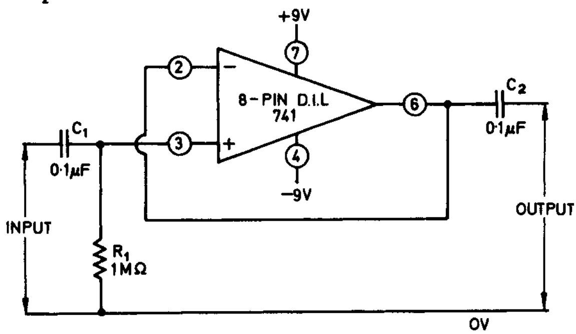


Fig. 1.12b. A.C. voltage follower

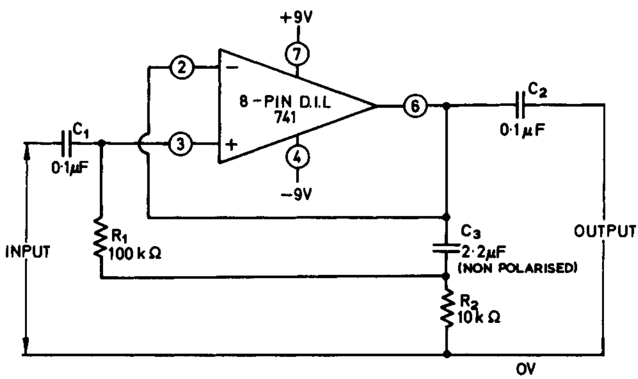


Fig. 1.12c. Very high input impedance a.c. voltage follower

The d.c. voltage follower can be adapted for a.c. use by wiring blocking capacitors in series with its input and output terminals and by d.c-coupling the non-inverting terminal of the op-amp to the zero volts line via a suitable resistor, as shown by R_1 in Fig. 1.12b. R_1 should have a value less than a couple of megohms, and restricts the available input impedance of the voltage follower.

If a very high input-impedance a.c. voltage follower is needed, the circuit of Fig. 1.12c can be used. Here, R_1 is bootstrapped from the output of the op-amp, and its apparent impedance is greatly increased. This circuit has a typical input impedance of hundreds of megohms.

The 741 op-amp is capable of providing output currents up to about 5mA, and this is consequently the current-driving limit of the three voltage follower circuits that we have looked at so far. The current-driving capabilities of the circuits can readily be increased by

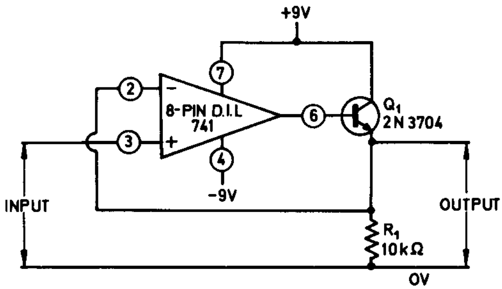


Fig. 1.13a. Unidirectional d.c. voltage follower with boosted output (variable from 0V to + 8V at 50mA)

wiring simple or complementary emitter follower booster stages between the op-amp output terminals and the outputs of the actual circuits, as shown in Figs. 1.13a and 1.13b respectively. Note in each

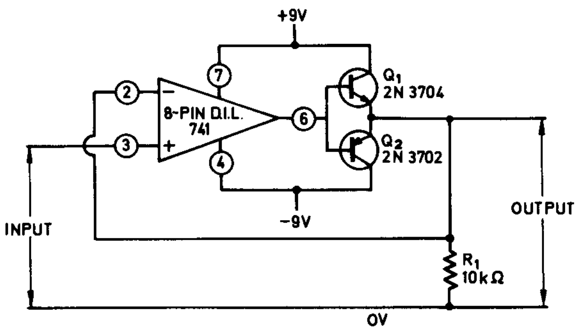


Fig. 1.13b. Bidirectional d.c. voltage follower with boosted output (variable from 0V to ± 8V at 50 mA)

case that the base-emitter junction(s) of the output transistor(s) are included in the negative feedback loop of the circuit. Consequently, the 600mV knee voltage of each junction is effectively reduced by a factor equal to the open-loop gain of the op-amp, so the junctions do not adversely effect the voltage-following characteristics of either circuit.

The Fig. 1.13a circuit is able to source current only, and can be regarded as a unidirectional, positive-going, d.c. voltage follower. The Fig. 1.13b circuit can both source (supply) and sink (absorb) output currents, and thus gives bidirectional follower action. Each circuit has a current-driving capacity of about 50mA. This figure is dictated by the limited power rating of the specified output transistors. The drive capability can be increased by using alternative transistors.

Miscellaneous amplifier projects

Figs. 1.14 to 1.22 show a miscellaneous assortment of 741 amplifier projects, ranging from d.c. adding circuits to frequency-selective amplifiers.

Fig. 1.14 shows the circuit of a unity gain inverting d.c. adder, which gives an output voltage that is equal to the sum of the three

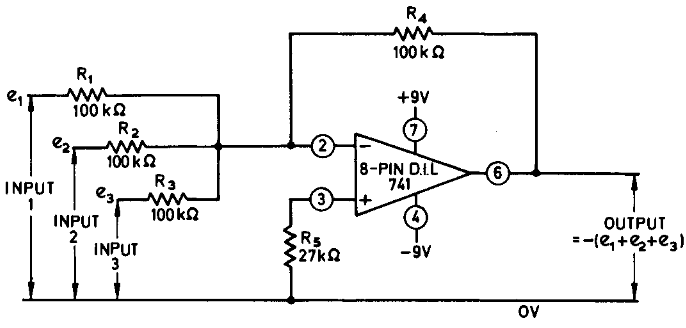


Fig. 1.14. Unity gain inverting d.c. adder or audio mixer'

input voltages. Here, input resistors R_1 to R_3 and feedback resistor R_4 each have the same value, and the circuit thus acts as a unity-gain inverting d.c. amplifier between each input terminal and the output. Since the current flowing in each input resistor also flows in feedback resistor R_4, the total current flowing in R_4 is equal to the sum of the input currents, and the output voltage is equal to the sum of the input voltages. The circuit is shown with only three input connections, but in fact can be provided with any number of input terminals. The circuit can be made to function as a so-called 'audio mixer' by wiring blocking capacitors in series with each input terminal and with the output terminal.

Fig. 1.15 shows how two unity-gain inverting d.c. amplifiers can be wired in series to make a precision unity-gain balanced d.c. phase-splitter. The output of the first amplifier is an inverted version of the input signal, and the output of the second amplifier is a non-inverted version.

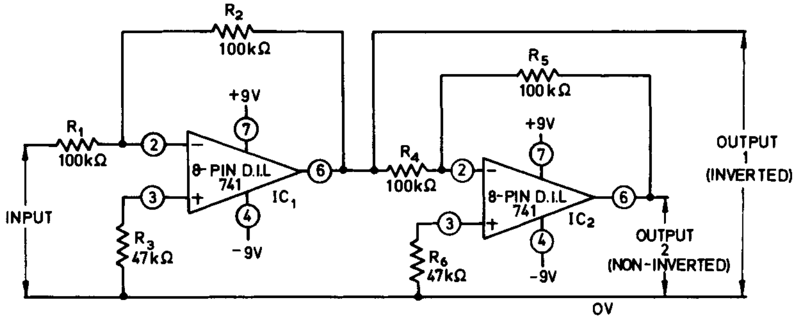


Fig. 1.15. Unity gain balanced d.c. phase-splitter

Fig. 1.16 shows how the 741 op-amp can be used as a unity-gain differential d.c. amplifier. The output of this circuit is equal to the difference between the two input signals or voltages, or to $e_1 - e_2$.

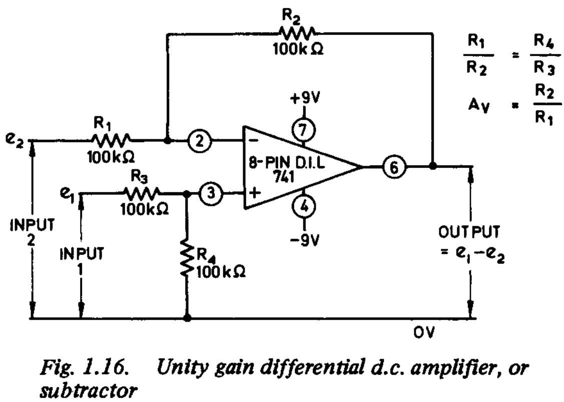
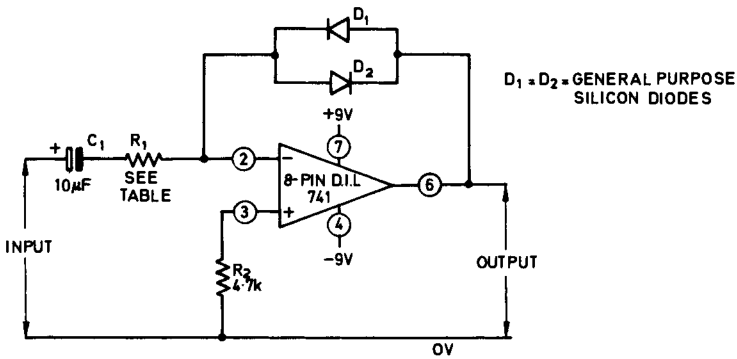


Fig. 1.16. Unity gain differential d.c. amplifier, or subtractor

Thus, the circuit can also be used as a subtractor. In this type of circuit the component values are chosen such that $R_1/R_2 = R_4/R_3$, in which case the voltage gain $A_V = R_2/R_1$. The circuit can thus be made to give voltage gain if required.

Fig. 1.17 shows how the 741 can be made to act as a non-linear (semi-log) a.c. voltage amplifier by using a couple of ordinary silicon



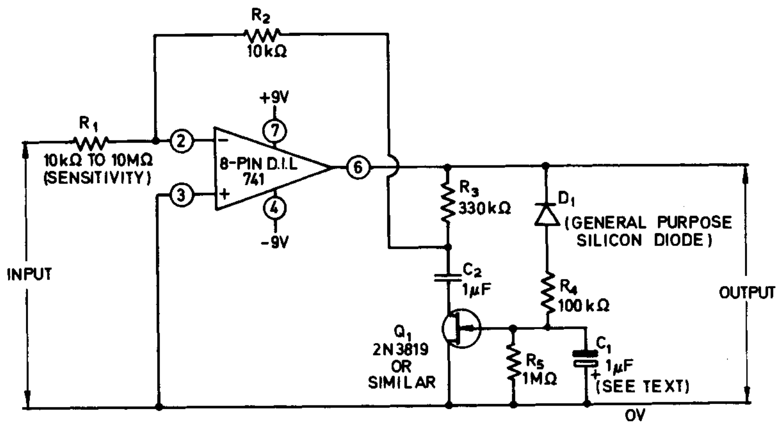
INPUT VOLTS (R.M.S)	R_1 = 1kΩ		R_1 = 10kΩ	
	V_{OUT} (R.M.S)	GAIN	V_{OUT} (R.M.S)	GAIN
1 mV	110 mV	X 110	21 mV	X 21
10 mV	330 mV	X33	170 mV	X 17
100 mV	450 mV	X4·5	360 mV	X 3·6
1V	560 mV	X0·56	470 mV	X 0·47
10V	600 mV	X0·07	560 mV	X0·056

Fig. 1.17. Circuit and performance table of non-linear (semi-log) a.c. voltage amplifier

diodes as feedback elements. The voltage gain of the circuit depends on the magnitude of applied input signal, and is high when input signals are low, and is low when input signals are high. The measured performance of the circuit is shown in the table, and can be varied by using alternative R_1 values.

Fig. 1.18 shows how the 741 can be used together with a junction-type field-effect transistor (j.f.e.t.) to make a so-called constant-volume amplifier. The action of this type of circuit is such that its peak output voltage is held sensibly constant, without distortion, over a wide range of input signal levels, and this particular circuit gives a sensibly constant output over a 30dB range of input signal levels. The measured performance of the circuit is shown in the table. C_1 determines the response time of the amplifier, and may be altered to satisfy individual needs.

The action of the Fig. 1.18 circuit relies on the fact that the j.f.e.t. can act as a voltage-controlled resistance which appears as a low value when zero bias is applied to its gate and as a high resistance when its gate is negatively biased. The j.f.e.t. and R_3 act as a gain-determining a.c. voltage divider (via C_2), and the bias to the j.f.e.t. gate is derived from the circuit output via the D_1–C_1 network. When the circuit output is low the j.f.e.t. appears as a low resistance, and the op-amp



V_{in} (R_1 = 10kΩ)	V_{in} (R_1 = 100kΩ)	V_{in} (R_1 = 1MΩ)	V_{in} (R_1 = 10MΩ)	V_{out}
50mV	500mV	5V	50V	2.85V
20mV	200mV	2V	20V	2.81V
10mV	100mV	1V	10V	2.79V
5mV	50mV	500mV	5V	2.60V
2mV	30mV	200mV	2V	2.03V
1mV	10mV	100mV	1V	1.48V
500µV	5mV	50mV	500mV	0.89V
200µV	2mV	20mV	200mV	0.40V
100µV	1mV	10mV	100mV	0.20V
50µV	500µV	5mV	50mV	0.10V

Fig. 1.18. Circuit and performance details of constant-volume amplifier

gives high voltage gain. When the circuit output is high the j.f.e.t. appears as a high resistance, and the op-amp gives low voltage gain. The output level of the circuit is thus held sensibly constant by negative feedback.

The 741 op-amp can be made to function as a frequency-selective amplifier by connecting frequency-sensitive networks into its feedback loops. Fig. 1.19 shows how a twin-T network can be connected to the op-amp so that it acts as a tuned (acceptor) amplifier, and Fig. 1.20 shows how the same twin-T network can be connected so that the

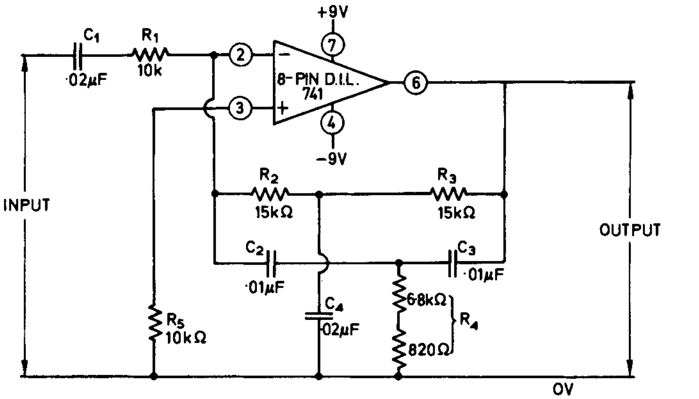


Fig. 1.19. 1kHz tuned (acceptor) amplifier (twin-T)

op-amp acts as a notch (rejector) filter. The values of the twin-T network are chosen such that $R_2 = R_3 = 2R_4$, and $C_2 = C_3 = C_4/2$, in which case its centre (tuned) frequency = $1/6.28\ R_2C_2$. With the component values shown, both circuits are tuned to approximately 1kHz.

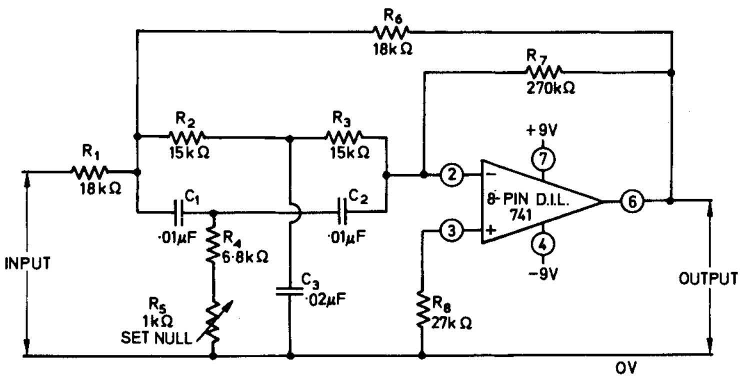


Fig. 1.20. 1kHz notch (rejector) filter

Finally, to complete this section of this chapter, Figs. 1.21 and 1.22 show the circuits of a couple of variable-frequency audio filters. The Fig. 1.21 circuit is that of a low-pass filter which covers the range 2.2kHz to 24kHz, and the Fig. 1.22 circuit is that of a high-pass filter which covers the range 235Hz to 2.8kHz. In each case, the circuit gives unity gain to signals beyond its cut-off frequency, and gives a second order response (a change of 12dB per octave) to signals within its range.

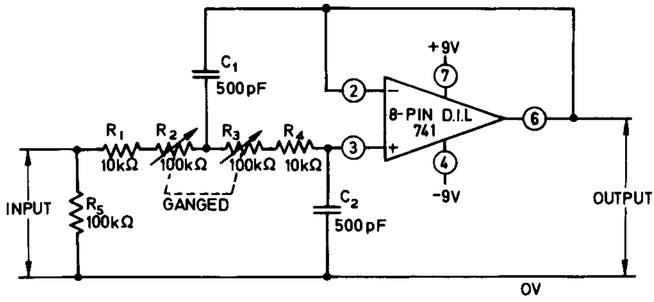


Fig. 1.21. Variable low-pass filter covering 2.2kHz to 24kHz

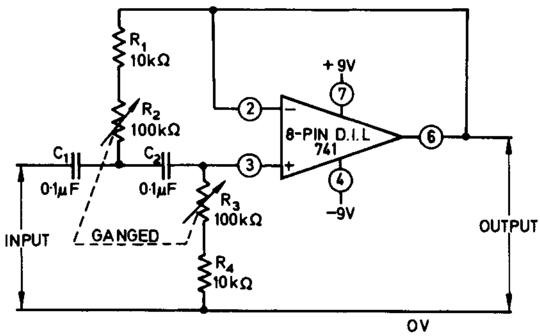


Fig. 1.22. Variable high-pass filter covering 235Hz to 2.8kHz

Instrumentation projects

Figs. 1.23 to 1.31 show a variety of instrumentation projects in which the 741 op-amp can be used. The circuits range from a simple voltage regulator to a linear-scale ohmmeter.

Fig. 1.23 shows the circuit of a simple variable-voltage power supply, which gives a stable output that is fully adjustable from 0V to 12V at currents up to a maximum of about 50mA. The operation of the circuit is quite simple. ZD_1 is a zener diode, and is energised from the positive supply line via R_1. A constant reference potential of 12V is developed across the zener diode, and is fed to variable potential divider R_2. The output of this divider is fully variable from 0V to 12V, and is fed to the non-inverting input of the op-amp. The op-amp is wired as a unity-gain voltage follower, with Q_1 connected as an emitter follower current-booster stage in series with its output. Thus, the output voltage of the circuit follows the voltage set at the op-amp input via R_2, and is fully variable from 0V to 12V. Note that the circuit uses an 18V positive supply and a 9V negative supply. Also note that the voltage range of

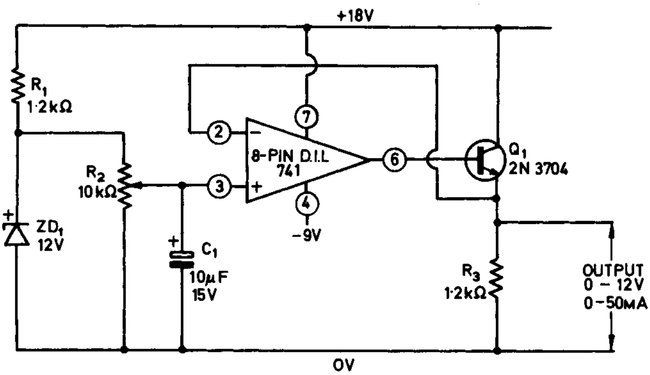


Fig. 1.23. Simple variable voltage supply

the above circuit can be increased by using higher zener and unregulated supply voltages, and that its current capacity can be increased by using one or more power transistors in place of Q_1.

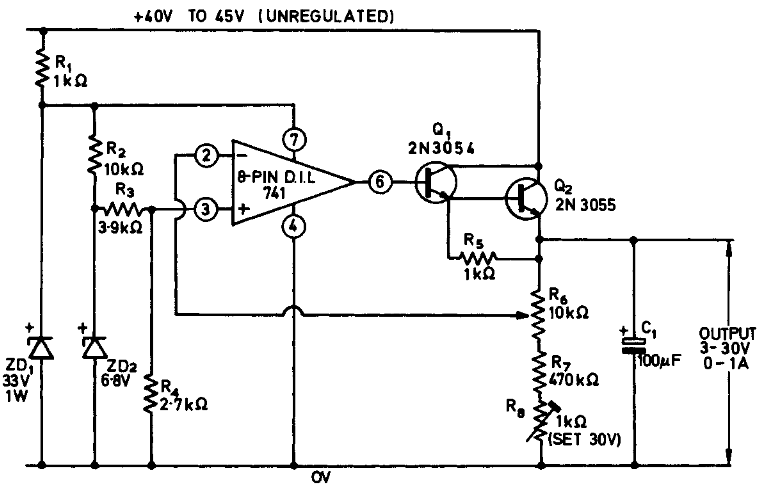


Fig. 1.24. 3V to 30V, 0 to 1A stabilised p.s.u.

Fig. 1.24 shows how the 741 op-amp can be used as the basis of a stabilised power supply unit (p.s.u.) that covers the range 3V to 30V at currents up to 1A. Here, the voltage supply to the op-amp is stabilised at 33V via ZD_1, and a highly temperature-stable reference of 3V is fed to the input of the op-amp via ZD_2. The op-amp and output transistors Q_1–Q_2 are wired as a variable-gain non-inverting d.c. amplifier, with gain variable from unity to x10 via R_6, and the output voltage is thus fully variable from 3V to 30V via R_6. The output voltage is fully stabilised by negative feedback.

Fig. 1.25 shows how overload protection can be applied to the above circuit. Here, current-sensing resistor R_9 is wired in series with the output of the regulator, and cut-out transistor Q_3 is driven from this resistor and is wired so that its base-collector junction as able to short

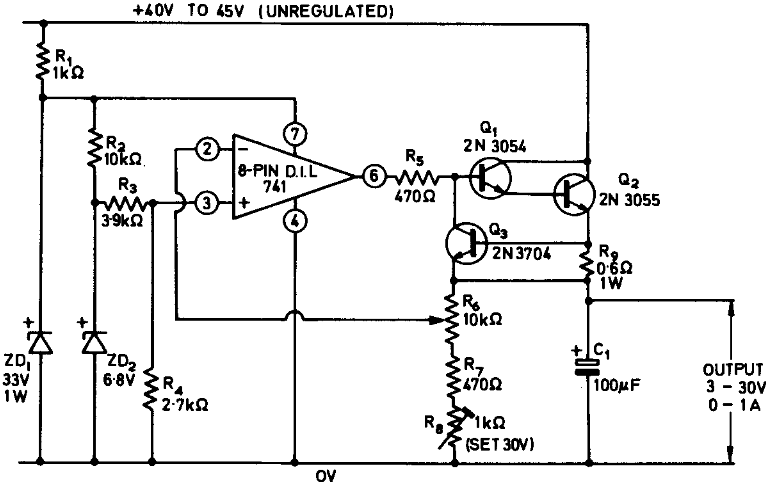


Fig. 1.25. 3V to 30V stabilised p.s.u. with overload protection

the base-emitter junction of the Q_1–Q_2 output transistor stage. Normally, Q_3 is inoperative, and has no effect on the circuit, but when p.s.u. output currents exceed 1A a potential in excess of 600mV is developed across R_9 and biases Q_3 on, thus causing Q_3 to shunt the base-emitter junction of the Q_1–Q_2 output stage and hence reducing the output current. Heavy negative feedback takes place in this action, and the output current is automatically limited to 1A, even under short-circuit conditions.

Fig. 1.26a shows how a 741 op-amp can be used in conjunction with a couple of silicon diodes as a precision half-wave rectifier. Conventional diodes act as imperfect rectifiers of low-level a.c. signals, because they do not begin to conduct significantly until the applied signal voltage exceeds a 'knee' value of about 600mV. When diodes are wired into the negative feedback loop of the circuit as shown the 'knee' voltage is effectively reduced by a factor equal to the open-loop gain of the op-amp, and the circuit thus acts like a near-perfect rectifier.

The overall voltage gain of the Fig. 1.26a circuit is dictated by the ratios of R_1 and R_2 to R_3, as in the case of a conventional inverting amplifier, and this circuit thus gives a gain of unity. The circuit can be made to act as a precision half-wave a.c./d.c. converter by designing it

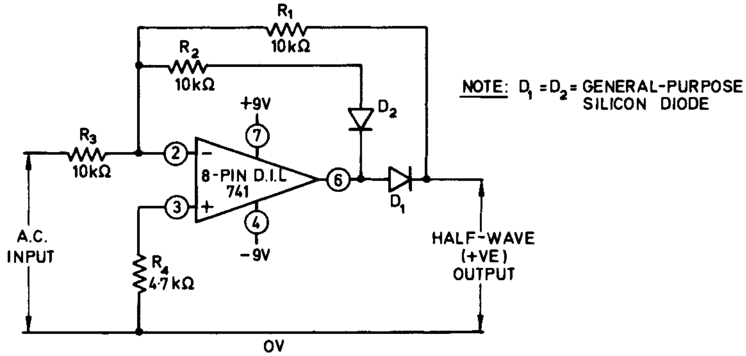


Fig. 1.26a. Precision unity gain half-wave rectifier

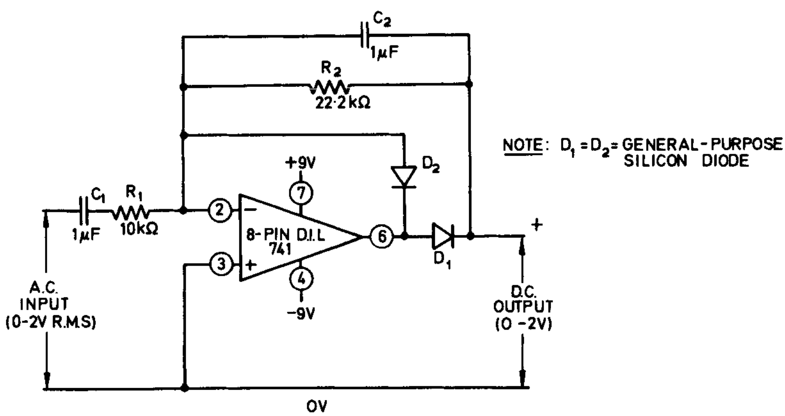
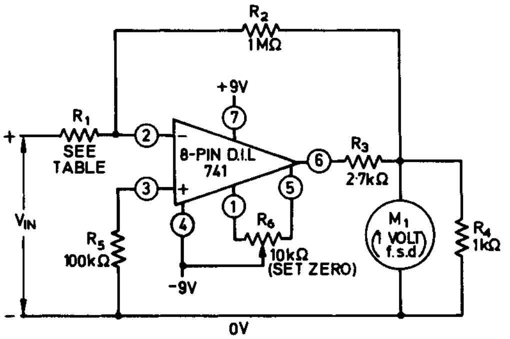


Fig. 1.26b. Precision half-wave a.c./d.c. converter

to give a voltage gain of 2.22 to give form-factor correction, and by integrating its rectifier output, as shown in Fig. 1.26b. Note that each of the Fig. 1.26 circuits has a high output impedance, and the outputs must not be fed into loads having impedances less than about 1MΩ.

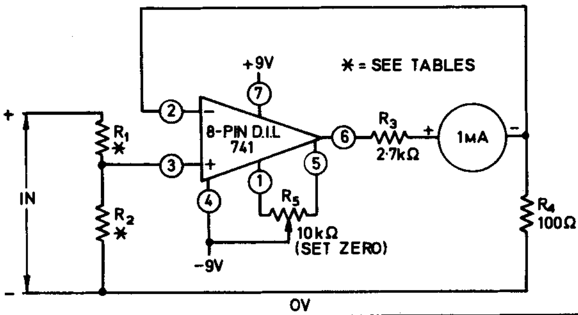
Fig. 1.27 shows how the op-amp can be used as a high-performance d.c. voltmeter converter, which can be used to convert any 1V f.s.d.

meter with a sensitivity better than 1kΩ/V into a voltmeter that can read any value in the range 1mV to 10V f.s.d. at a sensitivity of 1MΩ/V. The voltage range is determined by the R_1 value, and the table shows some suitable values for common voltage ranges.



$V_{f.s.d}$	R_1
10V	10MΩ
1V	1MΩ
100mV	100kΩ
10mV	10kΩ
1mV	1kΩ

Fig. 1.27. High performance d.c. voltmeter converter



VOLTMETER		
f.s.d	R_1	R_2
1000V	10MΩ	1kΩ
100V	10MΩ	10kΩ
10V	10MΩ	100kΩ
1V	900kΩ	100kΩ
100mV	0Ω	100kΩ

CURRENT METER		
f.s.d	R_1	R_2
1A	0Ω	0·1Ω
100mA	0Ω	1Ω
10mA	0Ω	10Ω
1mA	0Ω	100Ω
100μA	0Ω	1kΩ
10μA	0Ω	10kΩ
1μA	0Ω	100kΩ

Fig. 1.28. Simple d.c. voltage or current meter

Fig. 1.28 shows a simple circuit that can be used to convert a 1mA f.s.d. meter into a d.c. voltmeter with any f.s.d. value in the range 100mV to 1000V, or into a d.c. current meter with any f.s.d. value in the range 1μA to 1A. Suitable component values for different ranges are shown in the tables.

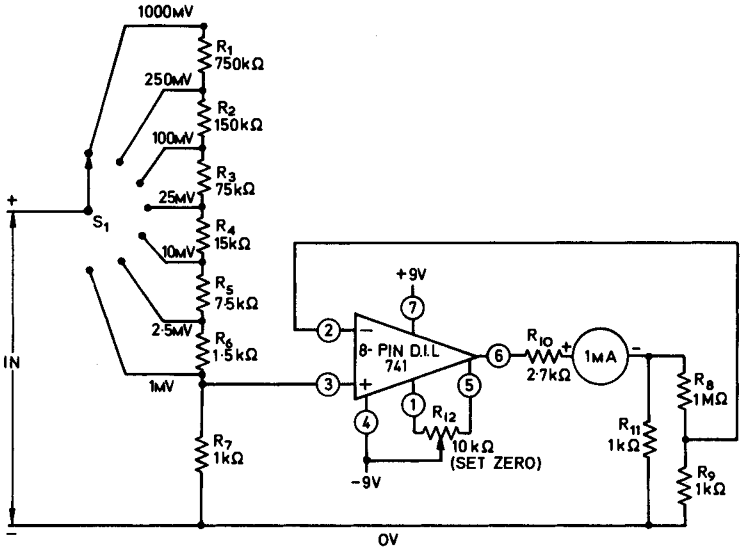
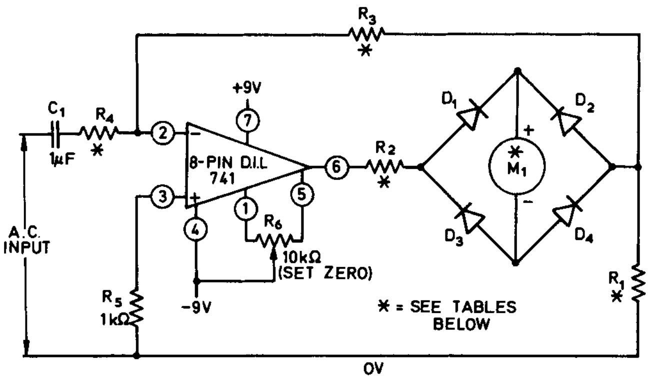


Fig. 1.29. Precision d.c. millivoltmeter



NOTE: D_1 TO D_4 = GENERAL-PURPOSE GERMANIUM DIODES

M_1	R_1	R_2
100µA	9kΩ	27kΩ
500µA	1·8kΩ	5·6kΩ
1MA	900Ω	2·7kΩ
2·5MA	360Ω	1·5kΩ
5MA	180Ω	470Ω

VALUES FOR USE WITH DIFFERENT METER MOVEMENTS

$V_{f.s.d}$	R_4	R_3
1000V	10MΩ	10kΩ
100V	10MΩ	100kΩ
10V	10MΩ	1MΩ
1V	1MΩ	1MΩ
100MV	100kΩ	1MΩ
10MV	10kΩ	1MΩ
1MV	1kΩ	1MΩ

DIFFERENT f.s.d VOLTAGE SENSITIVTIES

Fig. 1.30. Precision a.c. volt/millivolt meter

Fig. 1.29 shows the circuit of a precision d.c. millivoltmeter, which uses a 1mA f.s.d. meter to read f.s.d. voltages from 1mV to 1000mV in seven switch-selected ranges.

Fig. 1.30 shows the basic circuit of a precision a.c. volt or millivolt meter. This circuit can be used with any moving-coil meter with a full scale current value in the range 100μA to 5mA, and can be made to give any full scale a.c. voltage reading in the range 1mV to 1000V. The tables show the alternative values of R_1 and R_2 that must be used to satisfy different basic meter sensitivities, and the values of R_3 and R_4 that must be used for different f.s.d. voltage sensitivities.

Finally, to conclude this section of this chapter, Fig. 1.31 shows how the 741 op-amp can be used in conjunction with a 1mA f.s.d.

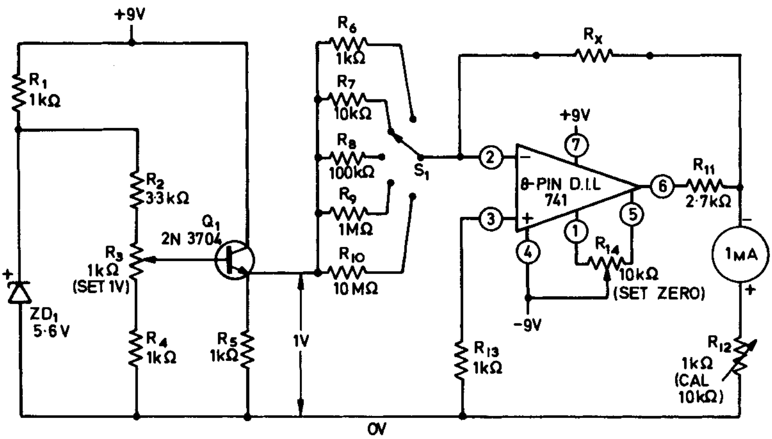


Fig. 1.31. Linear scale ohmmeter

meter to make a linear-scale ohmmeter that has five decade ranges from 1kΩ to 10MΩ. The circuit is divided into two parts, and consists of a voltage generator that is used to generate a standard test voltage, and a readout unit which indicates the value of the resistor under test.

The voltage generator section of the circuit comprises zener diode ZD_1, transistor Q_1, and resistors R_1 to R_5. The action of these components is such that a stable reference potential of 1V is developed across R_5, but is adjustable over a limited range via R_3. This voltage is fed to the input of the op-amp readout unit. The op-amp is wired as an inverting d.c. amplifier, with the 1mA meter and R_{12} forming a 1V f.s.d. meter across its output, and with the op-amp gain determined by the values of ranging resistors R_6 to R_{10} and by negative feedback resistor R_X. Since the input to the amplifier is fixed at 1V, the output voltage reading of the meter is directly proportional to the value of R_X, and equals full scale when R_X and the ranging resistor values are equal. Consequently, the circuit functions as a linear-scale ohmmeter.

The procedure for initially calibrating the Fig. 1.31 circuit is as follows. First, switch the unit to the 10kΩ range and fix an accurate 10kΩ resistor in the R_X position. Now adjust R_3 to give an accurate 1V across R_5, and then adjust R_{12} to give a precise full scale reading on the meter. All adjustments are then complete, and the circuit is ready for use.

Miscellaneous 741 projects

The type 741 op-amp can be used as the basis of a vast range of miscellaneous projects, including oscillators and sensing circuits. Four such projects are described in this final section of this chapter.

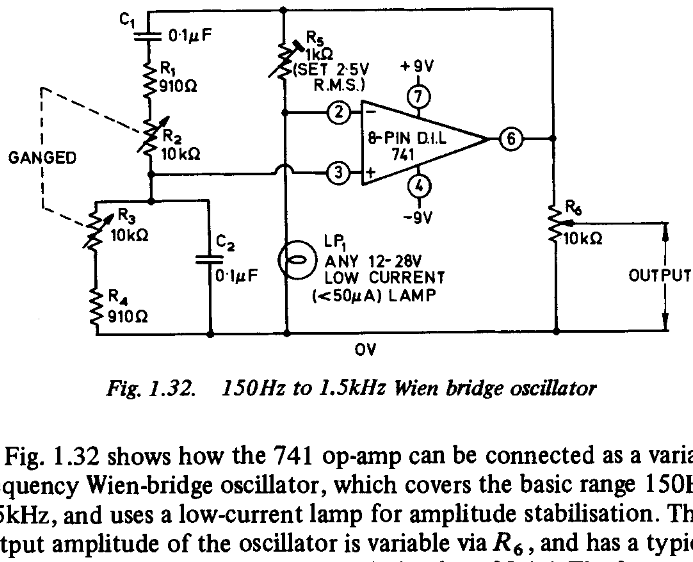


Fig. 1.32. 150Hz to 1.5kHz Wien bridge oscillator

Fig. 1.32 shows how the 741 op-amp can be connected as a variable-frequency Wien-bridge oscillator, which covers the basic range 150Hz to 1.5kHz, and uses a low-current lamp for amplitude stabilisation. The output amplitude of the oscillator is variable via R_6, and has a typical maximum value of 2.5V r.m.s. and a t.h.d. value of 0.1%. The frequency range of the circuit is inversely proportional to the C_1–C_2 values. The circuit can give a useful performance up to a maximum frequency of about 25kHz.

Fig. 1.33 shows how either a 741 or a 709 op-amp can be connected as a simple variable-frequency square-wave generator that covers the range 500Hz to 5kHz via a single variable resistor. The circuit produces a good symmetrical waveform. The frequency of oscillation is inversely proportional to the C_1 value, and can be reduced by increasing the C_1 value, or vice versa. The amplitude of the square wave output signal can

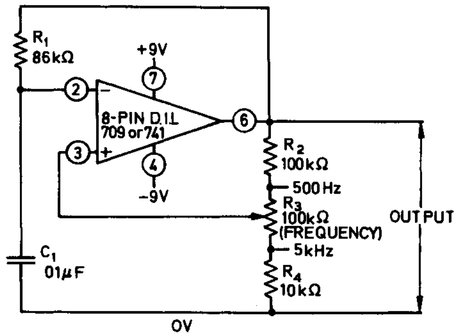


Fig. 1.33. Simple 500Hz to 5kHz square wave generator

be made variable, if required, by wiring a 10kΩ variable potential divider across the output terminals of the circuit and taking the output from between the pot slider and the zero volts line.

Finally, Figs. 1.34 and 1.35 show a couple of useful ways of using the 741 op-amp in the open-loop differential voltage comparator mode.

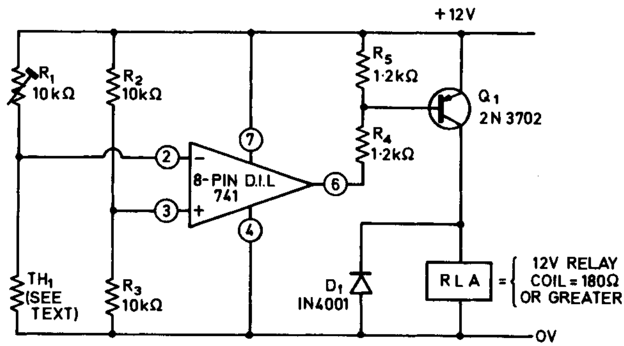


Fig. 1.34. Precision frost or under-temperature switch can be made to act as a fire or over-temperature switch by transposing R_1 and TH_1 positions

In each case, the circuits are powered from single-ended 12V supplies, and have a fixed half-supply reference voltage applied to the non-inverting op-amp terminal via the R_2–R_3 potential divider and have a variable voltage applied to the inverting op-amp terminal via a variable potential divider. The circuit action is such that the op-amp output is driven to negative saturation and the relay is driven on when the variable input voltage is greater than the reference voltage, and the op-amp output is driven to positive saturation and the relay is cut off when the variable input voltage is less than the reference voltage.

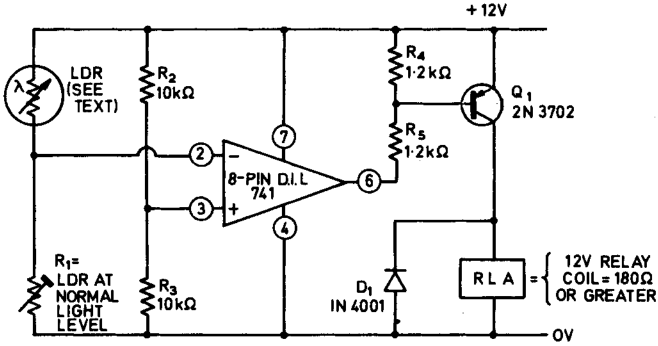


Fig. 1.35. Precision light-activated switch can be made to act as a dark-activated switch by transposing R_1 and LDR positions

The Fig. 1.34 circuit is that of a precision frost or under-temperature switch, which drives the relay on when the temperature sensed by thermistor TH_1 falls below a value pre-set via R_1. The circuit action can be reversed, so that it operates as a fire or over-temperature switch, by simple transposing the R_1 and TH_1 positions. In either case, TH_1 can be any negative-temperature-coefficient thermistor that presents a resistance in the range 900Ω to 9kΩ at the required trip temperature.

The Fig. 1.35 circuit is that of a precision light-activated switch, which turns the relay on when the illumination level sensed by light-dependent resistor *LDR* exceeds a value preset by R_1. The circuit action can be reversed, so that the relay turns on when the illumination falls below a preset level, by simply transposing the R_1 and *LDR* positions. In either case, the *LDR* can be any cadmium sulphide photocell that presents a resistance in the range 900Ω to 9kΩ at the desired switch-on level.

Footnote: In this chapter we have briefly described some thirty-five practical 741 op-amp projects. Readers seeking more information on these particular circuits, or looking for additional op-amp application information, will find such information in the Newnes-Butterworth publication *110 Operational Amplifier Projects*, by R. M. Marston.

CHAPTER 2

TYPE 555 TIMER APPLICATIONS

The type 555 timer is a highly versatile low-cost integrated circuit that is specifically designed for precision timing applications, but which can also be used in a variety of monostable multivibrator, astable multivibrator, and Schmitt trigger applications. The device was originally introduced by Signetics, but is now available under the '555' designation from most other i.c. manufacturers.

The 555 timer i.c. has many attractive features. It can operate from any supply voltages in the range 4.5V to 16V. Its output can source (supply) or sink (absorb) any load current up to a maximum of 200mA, so can directly drive loads such as relays, l.e.d.s, low-power lamps, and high impedance speakers. When used in the 'timing' mode, the i.c. can readily produce accurate timing periods that can be varied from a few microseconds to several hundred seconds via a single R–C network. Timing periods are virtually independent of actual supply rail voltage, and have a temperature coefficient of only 0.005% per degree C. Timing periods can be started via a *trigger* command signal, and can be aborted by a *reset* command signal.

When used in the monostable mode, the i.c. produces output pulses with typical rise and fall times of a mere 100ns. The i.c. can be made to produce pulse-width modulated (p.w.m.) pulses in this mode by feeding fixed frequency clock pulses to the *trigger* terminal of the i.c., and by feeding the modulation signal to the *control voltage* terminal.

When used in the astable mode both the frequency and the duty cycle of the waveform can be accurately controlled with two external resistors and one capacitor. The output signals can be subjected to frequency sweep control, frequency modulation (f.m.), or pulse-position modulation (p.p.m.) by applying suitable modulation signals to the *control voltage* terminal of the i.c.

How it works

The type 555 timer i.c. is available under a variety of specific type numbers, but is generally referred to simply as a '555 timer'. The device is available in a number of packaging styles, including 8- and 14-pin dual-in-line (d.i.l.) and 8-pin TO-99 types. Throughout this chapter, all circuits are designed around the standard 8-pin d.i.l. versions of the device.

Fig. 2.1 shows the outline and pin notations of the standard 8-pin d.i.l. version of the i.c., and Fig. 2.2 shows the functional block diagram

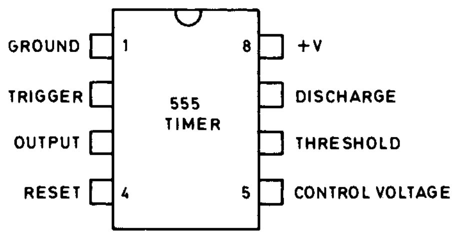


Fig. 2.1. Outline and pin notations of the standard 8-pin d.i.l. version of the 555 timer i.c.

of the same device (within the dotted lines), together with the connections for using the i.c. as a basic monostable generator. The following explanation of device operation assumes that the timer is used in the monostable configuration shown in Fig. 2.2.

The 555 timer houses 2 diodes, 15 resistors, and 23 transistors. These components are arranged in the form of one voltage-reference potential divider, two voltage-comparator op-amps, one R–S flip-flop, one low-power complementary output stage, and one slave transistor. The voltage-reference potential divider comprises three 5kΩ resistors in series, and is connected across the i.c. supply lines. Consequently, 2/3 V_{CC} appears at the junction of the upper two resistors of the potential divider, and is fed to one input terminal of the upper voltage-comparator op-amp, and 1/3 V_{CC} appears at the junction of the two lower resistors of the potential divider, and is fed to one input terminal of the lower voltage-comparator op-amp. The outputs of the two comparators control the R–S flip-flop, which in turn controls the states of the complementary output stage and the slave transistor. The state of the flip-flop can also be influenced by signals applied to the pin 4 *reset* terminal of the i.c.

When the monostable or timing circuit of Fig. 2.2 is in its quiescent state the pin 2 *trigger* terminal of the i.c. is held high via R_1. Under this condition Q_1 is driven to saturation and forms a short circuit across

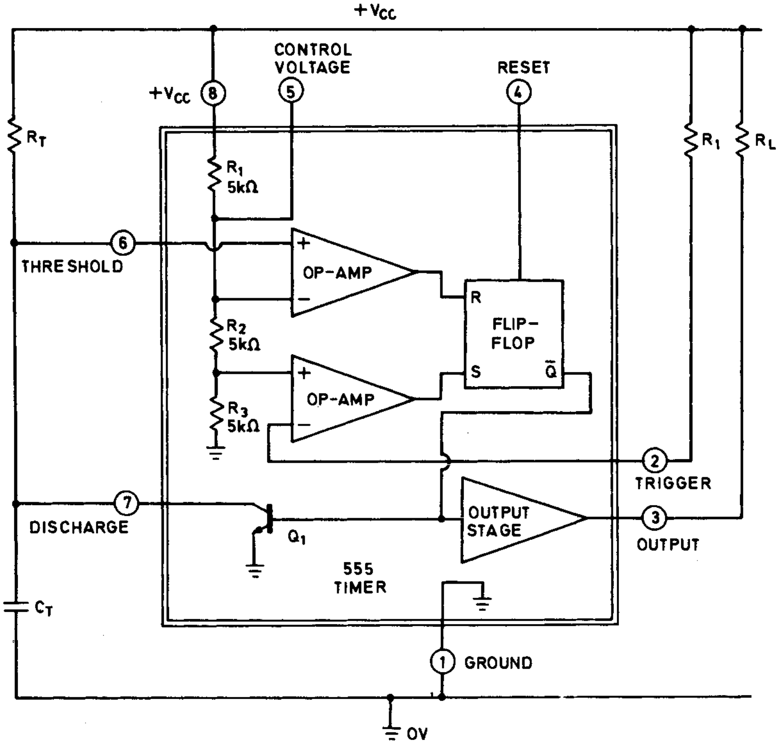


Fig. 2.2. Functional block diagram (within the square) of the 555 timer i.c., together with the connections for using the i.c. as a basic monostable generator or timer

external timing capacitor C_T, and the pin 3 output terminal of the i.c. is driven to the low state. The monostable action can be initiated by applying a negative-going trigger pulse to pin 2 of the i.c. As this pulse falls below the 1/3 V_{CC} reference value of the built-in potential divider the output of the lower voltage comparator op-amp changes state and causes the R–S flip-flop to switch over, which cuts off Q_1 and drives the pin 3 output of the i.c. to the high state.

As Q_1 cuts off it removes the short from timing capacitor C_T, so C_T starts to charge exponentially towards the supply rail voltage until eventually the C_T charge reaches 2/3 V_{CC}. At this point the upper voltage comparator op-amp changes state and switches the R–S flip-flop back to its original condition, so Q_1 turns on and rapidly discharges C_T, and simultaneously the pin 3 output of the i.c. reverts to its low state. The monostable operating sequence is then complete. Note that, once triggered, the circuit cannot respond to additional triggering until the timing sequence is complete, but that the sequence can be aborted at any time by feeding a negative-going pulse to pin 4 of the i.c.

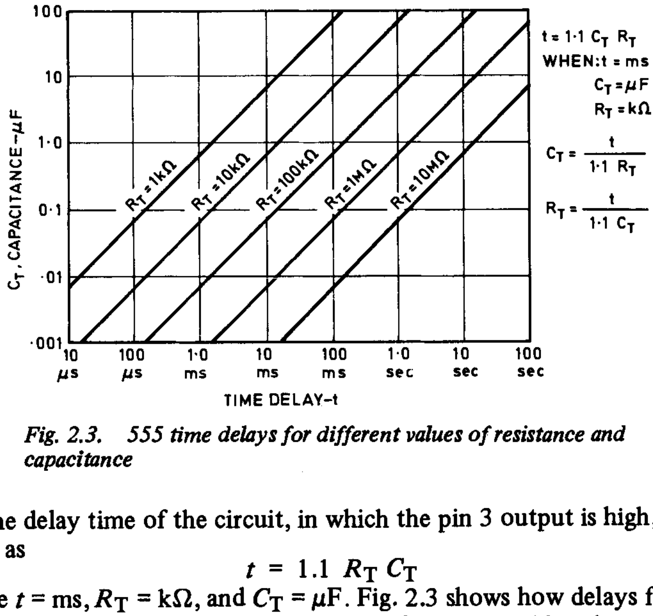


Fig. 2.3. 555 time delays for different values of resistance and capacitance

The delay time of the circuit, in which the pin 3 output is high, is given as

$$t = 1.1\ R_T\ C_T$$

where t = ms, R_T = kΩ, and C_T = μF. Fig. 2.3 shows how delays from 10μs to 100 seconds can be obtained by selecting suitable values of C_T and R_T in the range 0.001μF to 100μF and 1kΩ to 10MΩ. In practice, R_T should not be given a value less than 1kΩ or greater than 20MΩ, and capacitor C_T must always be a low-leakage component. Note that the timing period of the circuit is virtually independent of the circuit's actual supply voltage value, but that the period can be varied by applying a variable resistance or voltage between the ground and pin 5 *control voltage* terminals of the i.c. This facility enables the periods to be externally modulated or compensated.

The pin 3 output terminal of the i.c. is normally low, but switches high during the active monostable sequence. The output can either source or sink currents up to a maximum of 200mA, so external loads can be connected between pin 3 and either the positive supply rail or the ground rail, depending on the type of load operation that is required. The output switching rise and fall times are typically about 100 nanoseconds.

Practical timer circuits

Fig. 2.4 shows the practical circuit and waveforms of a simple manually triggered 50 second timer or pulse generator that gives a direct voltage output at pin 3. The output voltage is normally low, but goes high for the duration of the timing period. Optional components R_4 and l.e.d. (shown dotted) give a visual indication of the timer action. The circuit

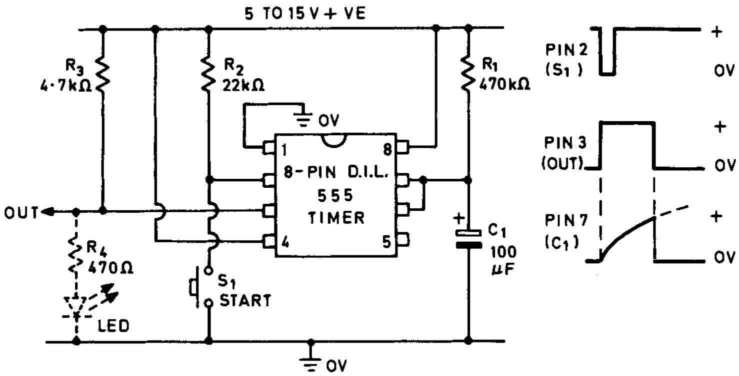


Fig. 2.4. Circuit and waveforms of simple manually-triggered 50 second timer or pulse generator

works in the same basic way as already described, except that the timing action is initiated by momentarily shorting pin 2 to ground via *start* switch S_1. Note from the circuit waveforms that a fixed-period output pulse is available at pin 3 and an exponential sawtooth with an identical period is available at pin 7. The sawtooth waveform has a high output impedance.

The basic timer circuit of Fig. 2.4 can be varied in a number of ways, as shown in Fig. 2.5. Here, the timing period is made variable between approximately 1.1 seconds and 110 seconds by replacing R_1 with a 10kΩ fixed resistor and a 1MΩ variable resistor in series. The period can be further varied, if required, by switch-selecting decade values of timing capacitance. The diagram also shows how the circuit can be provided with a *reset* facility, so that a timing period can be aborted at any time, by taking pin 4 to the positive supply rail via resistor R_5 and wiring *reset* switch S_2 between pin 4 and ground.

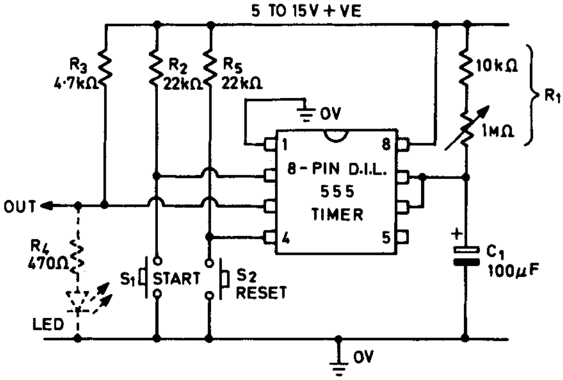


Fig. 2.5. Manually-triggered variable (1.1 to 110 seconds) timer with reset facility

The timing circuits of Figs. 2.4 and 2.5 can be used to drive non-inductive loads at currents up to 200mA directly. They can be used to drive inductive relay loads by using the basic connections shown in Fig. 2.6 or Fig. 2.7.

The Fig. 2.6 circuit is designed to apply a connection to a normally off external load for a preset period of 50 seconds when *start* switch S_1 is momentarily closed. The relay is normally off, but turns on for the 50 second period when the timing cycle is initiated. Here, D_1 is used to damp the potentially harmful back e.m.f. of the relay coil as it switches off at the end of the timing cycle, and D_2 is wired in series with the relay coil to counteract the slight residual voltage that appears at pin 3 of the i.c. under the *off* condition and thus ensure that the relay turns fully off.

The Fig. 2.7 circuit is designed to break the connection to a normally on external load for a preset period of 50 seconds when *start* switch S_1 is momentarily closed. The relay is normally on, but turns off for the 50 second period when the timing cycle is initiated. Here, the relay coil is wired between pin 3 and the positive supply line, and diode D_1 is wired across the coil to damp the back e.m.f. that occurs at the moment that the relay switches off at the start of the timing cycle.

Note in Figs. 2.6 and 2.7, and all other relay-output circuits described here, that the relays used can be any 12 volt types that draw *on* currents of less than 200mA, e.g., that have coil resistances greater than 60Ω.

The basic relay-driving timer circuits of Figs. 2.6 and 2.7 can be adapted for use in a variety of useful applications. Some typical examples are shown in Figs. 2.8 to 2.11.

Fig. 2.8 shows the practical circuit of a relay-output general-purpose timer that covers 0.9 seconds to 100 seconds in two decade ranges. The

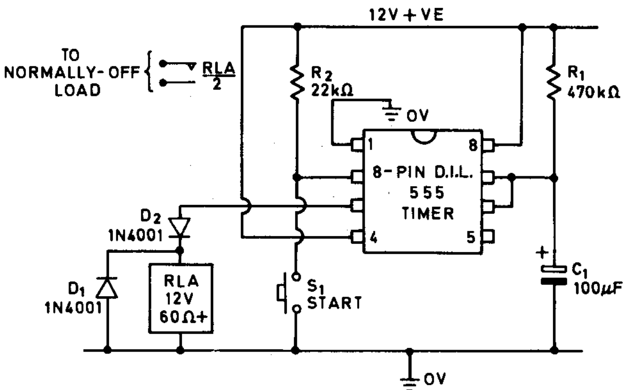


Fig. 2.6. Relay-output timer applies connection to normally-off load for preset period of 50 seconds when S_1 is momentarily operated

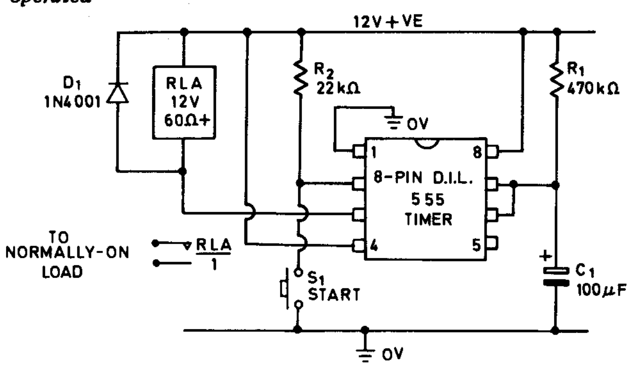


Fig. 2.7. Relay-output timer breaks connection to normally-on load for preset period of 50 seconds when S_1 is momentarily operated

circuit has a *reset* facility provided via S_2, so that timing periods can be aborted part way through a cycle if necessary. A noteworthy feature of this circuit is that the maximum timing periods of each decade range of the timer can be precisely preset via R_5 or R_6, which effectively shunt the built-in potential divider of the 555 timer i.c. and thus influence the timing periods. This facility enables the circuit to give precise timing periods even when wide-tolerance timing capacitors are used.

To set up the Fig. 2.8 circuit, first set R_1 to maximum value, set *range* switch S_3 to position 1, activate *start* switch S_1, and adjust R_5 to

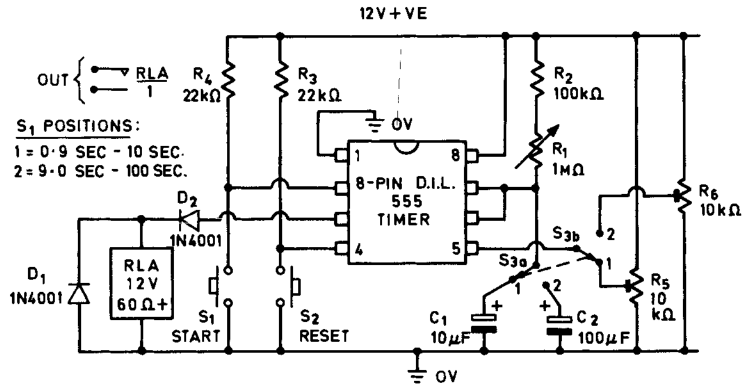


Fig. 2.8. Relay-output general-purpose timer covers 0.9 to 100 seconds in two decade ranges

give a timing period of precisely 10 seconds. Next, set S_3 to position 2, activate *start* switch S_1, and adjust R_6 to give a timing period of precisely 100 seconds. All adjustments are then complete, and the timer is ready for use.

Fig. 2.9 shows the practical circuit of an automatic delayed-turn-off headlight control system for automobiles. The idea here is that, if the headlights are deliberately left on when the car is parked, the time-delay circuit automatically switches them off about 50 seconds after the ignition switch is turned off. This facility enables the owner to use the car lights to illuminate his path for a preset time after parking as he leaves the garage or walks along a driveway, etc. The circuit does not interfere with normal headlight operation under actual driving conditions. The system works as follows.

When the ignition switch is turned to the *on* position current is fed to the coil of the relay via D_3 and the 12V supply rail, so the relay turns on and contacts RLA/1 close. As the contacts close they connect the 12V supply to the timer circuit and to the headlight switch. Thus, under this 'ignition on' condition the headlights operate in the normal way. Note that, since one side of C_2 is connected directly to the positive supply rail and the other side is taken to the positive rail via R_2, the capacitor is fully discharged under this condition.

The moment that the ignition switch is turned to the *off* position the D_3-derived current supply to the relay coil is broken, and simultaneously a negative-going trigger pulse is fed to pin 2 of the i.c. as the C_2–R_3 junction drops to ground volts and C_2 charges up. Now, relays are inherently slow-acting devices, so contacts RLA/1 do not open

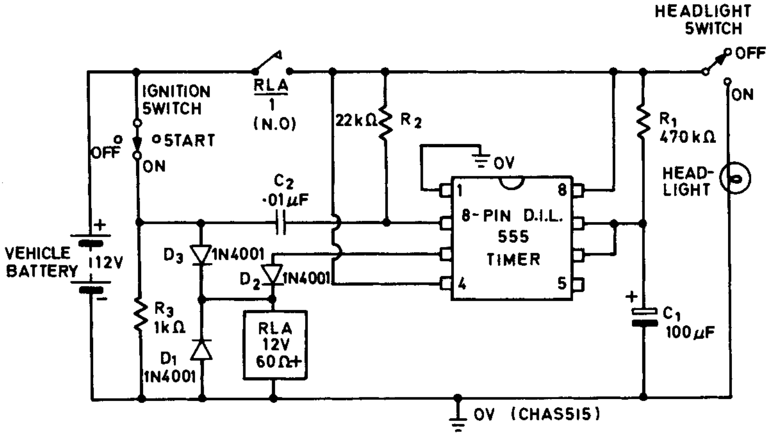


Fig. 2.9. Automatic delayed-turn-off headlight control system for automobiles

instantaneously as the ignition switch is turned off. Conversely, the 555 timer is a very fast triggering device, and the instant that the trigger pulse is generated via the turn-off action of the ignition switch a timing cycle is initiated and current is fed to the relay coil via output pin 3 of the i.c. as pin 3 goes high. Thus, the relay remains on for a preset period after the ignition switch is closed, and the positive supply rail remains connected to the headlight switch for the duration of this period. With the component values shown this period approximates 50 seconds.

At the end of the 50 second timing period, pin 3 of the i.c. switches to the low state and the relay turns off. As the relay turns off contacts RLA/1 open and remove the supply from the timer and the headlight switch, and the headlights turn off. The operating sequence is then complete.

Readers may care to note that the above system of operation is consistent with the practice adopted in many modern vehicles of feeding the headlight switch via the ignition switch, so that the headlights operate only when the ignition is turned on. On older types of vehicle, where headlight operation is independent of the ignition switch, a manually triggered delayed-turn-off headlight or spotlight control facility can be obtained by using the circuit shown in Fig. 2.10. The action of this circuit is such that, if the vehicle is parked with its light off, the lights turn on for a preset 50 second period as soon as a push-button *start* switch is momentarily closed, and at the end of this period turn off again automatically.

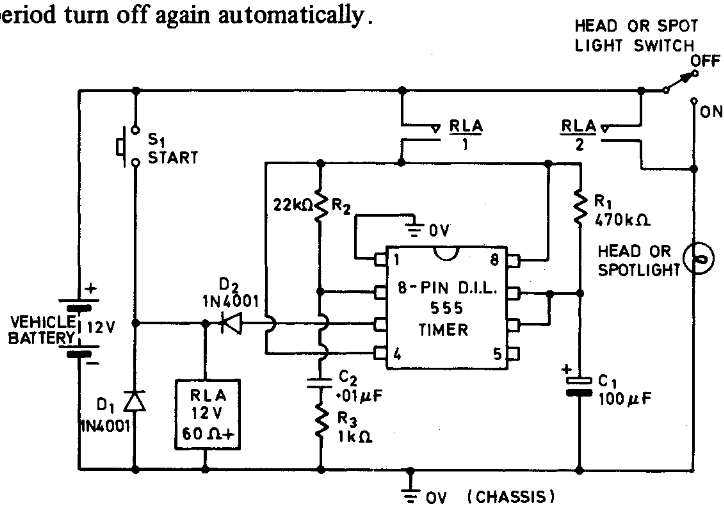


Fig. 2.10. Manually-triggered delayed-turn-off head or spot light control system for automobiles

The Fig. 2.10 circuit uses a relay with two sets of normally open relay contacts. The timing sequence is initiated by momentarily closing push-button switch S_1. Circuit operation is very simple. Normally, both S_1 and the relay contacts are open, so zero power is fed to the timer circuit and the lights are off. C_2 is discharged under this condition.

When S_1 is momentarily closed power is fed directly to the relay coil, and the relay turns on. As the relay turns on contacts RLA/2 close and apply power to the vehicle lights and contacts RLA/1 close and apply power to the timer circuit, but pin 2 of the i.c. is briefly tied to ground via C_2 and R_3 at this moment, so a negative trigger pulse is immediately fed to pin 2 of the i.c. and a timing cycle is initiated. Consequently, pin 3 of the i.c. switches high at the moment that the relay contacts close, and thus locks the relay into the *on* condition irrespective of the subsequent state of *start* switch S_1, so the lights remain on for the duration of the 50 second timing cycle. At the end of the timing cycle pin 3 of the i.c. switches to the low state, so the relay turns off and contacts RLA/1 and RLA/2 open, disconnecting power from the timing circuit and the lights. The operating sequence is then complete.

Finally, to conclude this 'timer circuits' section of the 555 story, Fig. 2.11 shows the circuit of a relay-output automatic porch light control unit that turns the porch lights on for a preset 50 second

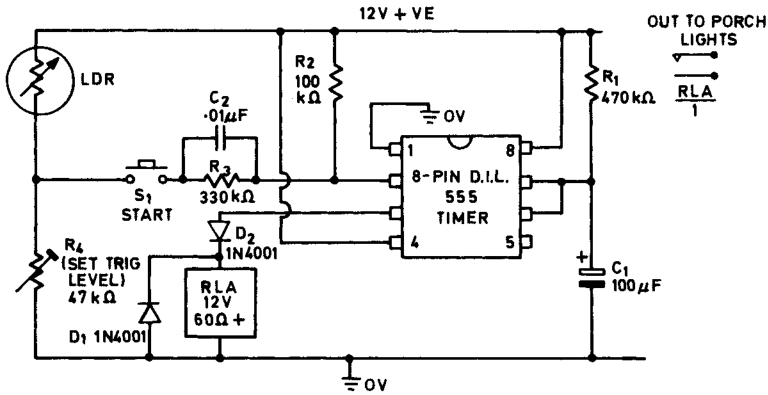


Fig. 2.11. Automatic porch light turns on for a preset period only when triggered at night

period only when suitably triggered at night time or under 'dark' conditions. The circuit is triggered via switch S_1, which may take the form of a microswitch activated by a porch gate or a pressure-pad switch activated by body weight and concealed under a porch mat or rug.

The operation of the Fig. 2.11 circuit relies on the fact that for correct timer operation the negative-going trigger pulse that is fed to pin 2 of the i.c. must fall below the internally-controlled 1/3 V_{CC} voltage value of the i.c. If the trigger pulse does not fall below this value, timing cycles can not be initiated by the trigger signal.

In the design, light-dependent resistor *LDR* and preset resistor R_4 are wired in series as a light-dependent potential divider. One side of switch S_1 is taken to the output of this potential divider, and the other side of the switch is taken to pin 2 of the i.c. via the C_2–R_3 combination. Under bright or daylight conditions the *LDR* acts as a low resistance, so a high voltage appears at the output of the potential divider. Consequently, the act of closing S_1 causes a voltage pulse much higher than 1/3 V_{CC} to be fed to pin 2 of the i.c., so the timer is not triggered via S_1 under the 'daylight' condition.

Conversely, the *LDR* acts as a high resistance under dark or 'night' conditions, so a low voltage appears at the output of the potential divider. Consequently, the act of closing S_1 causes a voltage pulse much lower than 1/3 V_{CC} to be fed to pin 2 of the i.c., so the timer circuit is triggered via S_1 under the 'night' condition.

In practice, the *LDR* can be any cadmium-sulphide photocell that presents a resistance in the range 1kΩ to 100kΩ under the required minimum 'dark' turn-on condition, and R_4 can be adjusted to preset the minimum 'dark' level at which the circuit will trigger. Note that the trigger signal is fed to pin 2 of the i.c. via the C_2–R_3 combination, which act as a trigger signal conditioning network that effectively isolates the d.c. component of the *LDR*–R_4 potential divider from the trigger pin of the i.c.

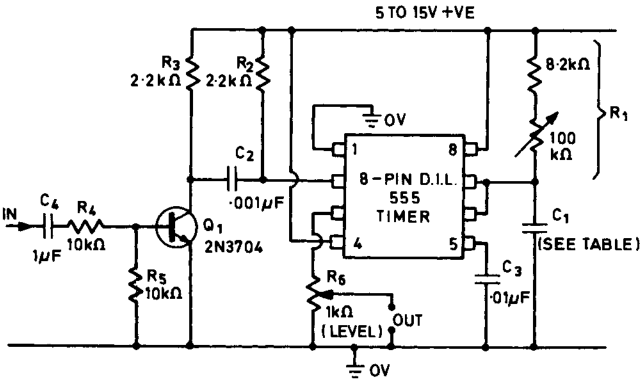
Monostable pulse generator circuits

All the 555 timer circuits that we have looked at so far act essentially as monostable multivibrators or pulse generators. The 555 timer i.c. can be used as a conventional electronically triggered monostable multivibrator or pulse generator by feeding suitable trigger signals to pin 2 of the i.c. and taking the pulse output signals from pin 3. The i.c. can be used to generate good output pulses with periods from 5μs to several hundred seconds. The maximum usable pulse repetition frequency is approximately 100kHz.

The trigger signal reaching pin 2 of the i.c. must be a carefully shaped negative-going pulse. Its amplitude must switch from an *off* value greater than 2/3 V_{CC} to an *on* value less than 1/3 V_{CC} (triggering actually occurs as pin 2 drops through the 1/3 V_{CC} value). The pulse must have a width greater than 100ns but less than that of the desired output pulse, so that the trigger pulse is removed by the time the monostable period terminates.

One way of obtaining a suitable trigger signal for the 555 monostable circuit is to convert the input signal to a good square wave that switches between ground volts and the full positive supply rail voltage, and then couple this square wave to pin 2 of the i.c. via a simple short time constant C–R differentiating network, which converts the leading or trailing edges of the square wave into suitable trigger pulses. Fig. 2.12a shows a practical circuit that uses this basic principle, but is intended for use only with input signals that are already of square or pulse form.

Here, transistor Q_1 converts the rectangular input signal into a signal that switches between the ground and positive voltage rails, and the resulting signal is fed to pin 2 of the i.c. via the C_2–R_2 differentiating network. The circuit can be used as an add-on pulse generator that is used in conjunction with an existing square or pulse generator. Variable-amplitude output pulses are available from pin 3 via variable potential divider R_6. The output pulse widths can be varied over more than a



C₁ VALUE	PULSE WIDTH RANGE
10μF	90 ms — 1·2 sec
1μF	9 ms — 120 ms
0·1μF	0·9ms — 12 ms
·01μF	90μs — 1·2 ms
·001μF	9μs — 120μs

Fig. 2.12a. Simple add-on pulse generator is triggered by rectangular input signals: circuit can be used at trigger frequencies up to 100kHz

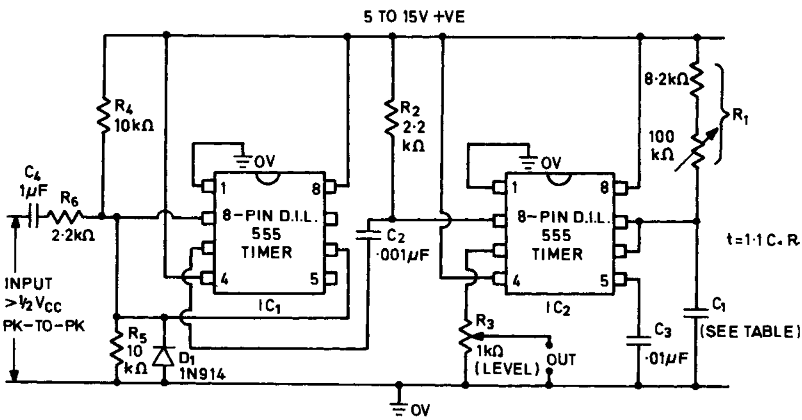


Fig. 2.12b. Improved add-on pulse generator is triggered by any input waveform

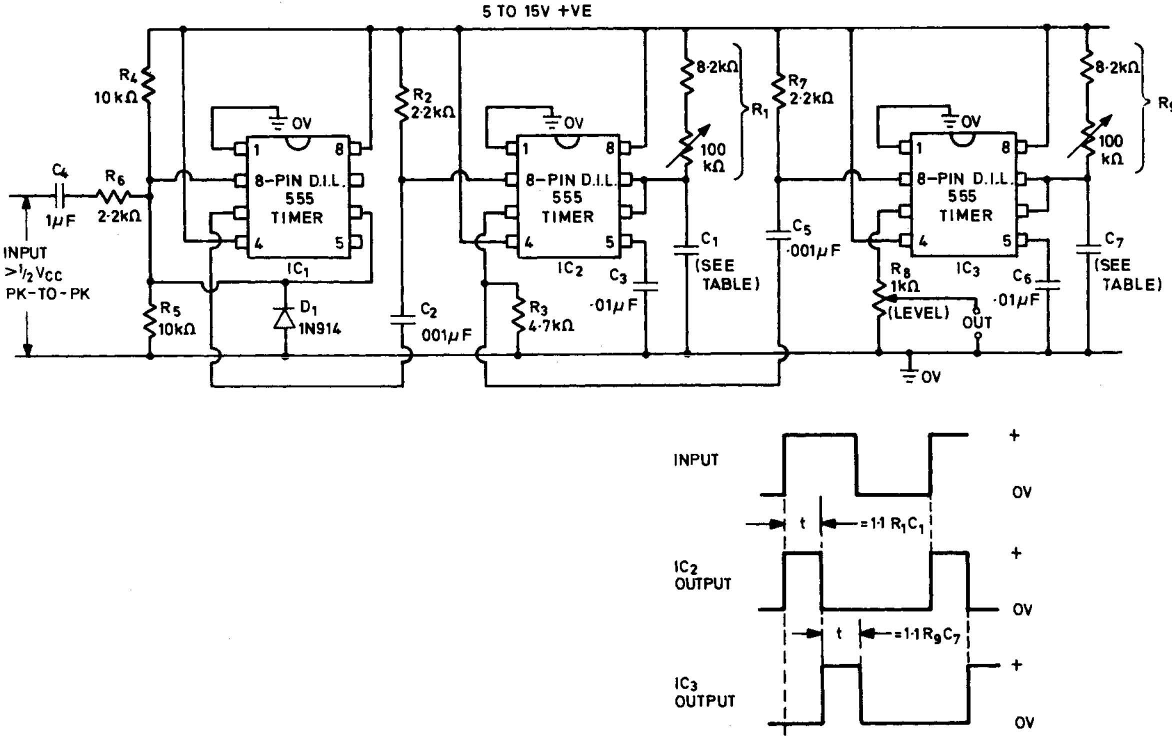


Fig. 2.13. Add-on delayed pulse generator is triggered by any input waveform. For C_1 (and C_7) values see table in Fig. 2.12

decade range via R_1, and can be switched in overlapping decade ranges by using the values of C_1 listed in the table. With the component values shown the pulse width is fully variable from 9μs to 1.2 seconds. Note that C_3 is used to decouple the pin 5 *control voltage* terminal of the i.c. and improve the circuit stability.

Fig. 1.12b shows how the above circuit can be modified so that it can be driven from any type of input waveform, including sine waves. Here, IC_1 is connected as a simple Schmitt trigger, which converts all input signals into rectangular output signals, and these rectangular signals are used to drive the IC_2 monostable circuit in the same way as described above. The Fig. 2.12b circuit can thus be used as an add-on pulse generator in conjunction with an existing waveform generator of any type that produces output signals with peak-to-peak amplitudes greater than 1/2 V_{CC}.

Note that the Fig. 2.12a and 2.12b circuits can readily be converted into self-contained pulse generators by simply building either of them into the same cabinet as a variable-frequency square wave generator, which can then be used to provide the necessary trigger signals.

Fig. 2.13 shows how two basic monostable pulse generators can be connected in series to make a delayed pulse generator, in which IC_1 is used as a Schmitt trigger and IC_2 controls the delay width and IC_3 determines the output pulse width. The final output pulse appears some delayed time after the initial application of the trigger signal. This circuit can again be made into a self-contained instrument by building it into the same cabinet as a simple square wave generator, which can be used to provide the necessary input drive signals.

Any number of basic monostable pulse generators can be wired in series to give a sequential form of operation. Fig. 2.14, for example, shows the circuit and waveforms of a three-stage sequential generator, which can be used to operate lamps or relays, etc., in a preprogrammed time sequence once an initial *start* command is given via push-button switch S_1. Note that the pin 4 *reset* terminal of all i.c.s are shorted together and positively biased via R_7, and that these terminals can be shorted to ground via *set* switch S_2. This *set* switch should be closed at the moment that power is first applied to the circuit, to ensure that none of the i.c.s are falsely triggered at this moment.

Finally, Fig. 2.15 shows how three or more monostable circuits can be connected in series in a continuous loop, with the output of the last monostable feeding back to the input of the first monostable, to form a 'chaser' circuit in which the sequential action repeats to infinity. This type of circuit can be used to drive lamp or l.e.d. displays, etc. Note that the circuit is again provided with the S_2 *set* facility, so that the circuit can be emptied at the moment that power is first applied.

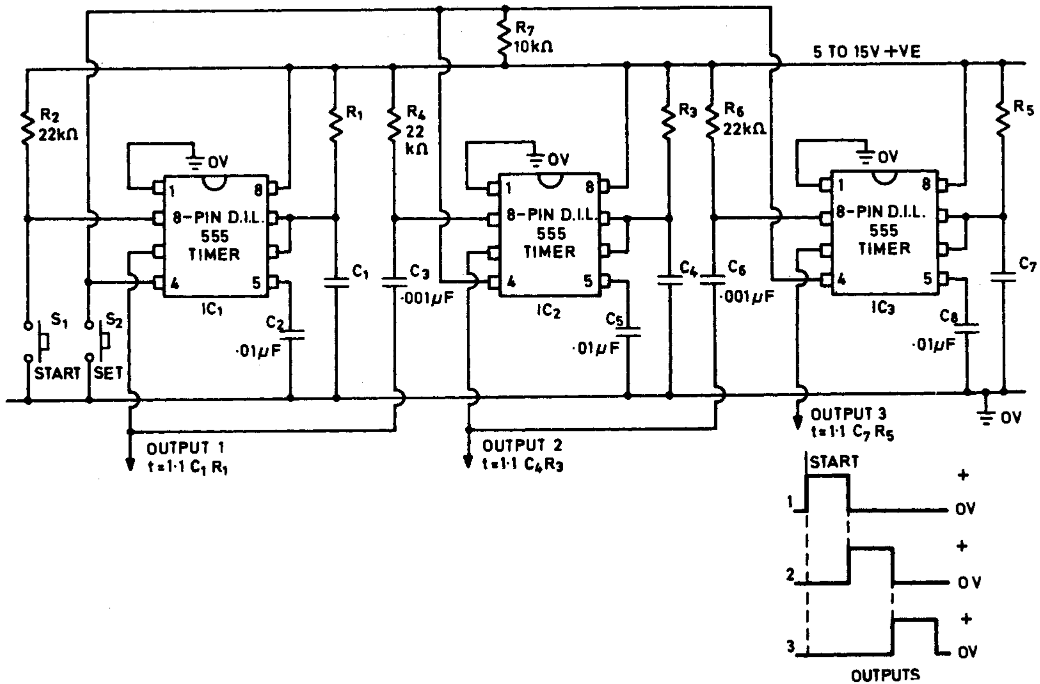


Fig. 2.14. Three-stage sequential timer or pulse generator

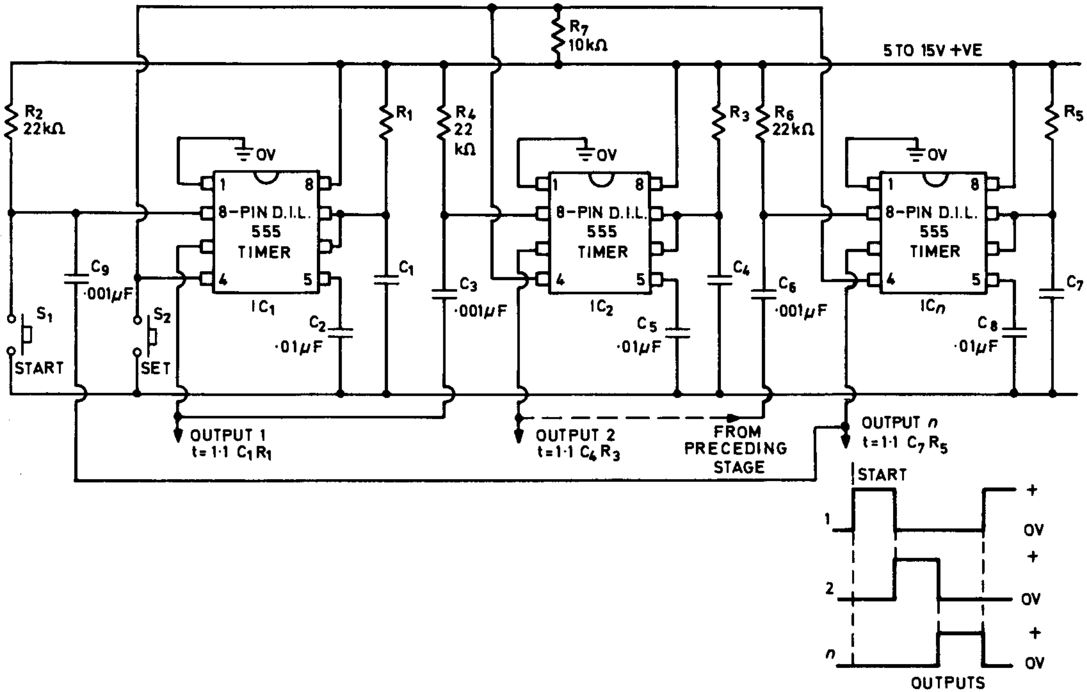
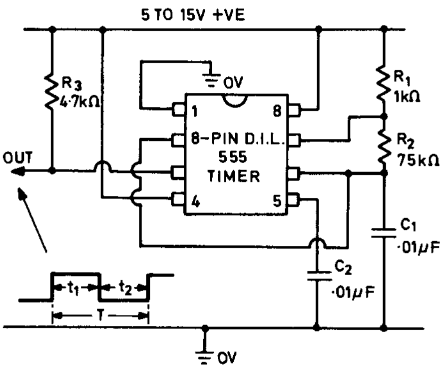


Fig. 2.15. *'Chaser' or regenerative sequential timer can have any number of stages*

Astable multivibrator circuits

The 555 timer i.c. can be used as an excellent astable multivibrator or square wave generator at operating frequencies up to 100kHz. The 555 astable has excellent frequency stability, has a low output impedance, and gives typical output rise and fall times of only 100ns. Other advantages of the circuit are that its duty cycle can readily be adjusted, and that the output signals can readily be subjected to frequency modulation via pin 5 of the i.c.



	IF $R_2 \gg R_1$:
$t_1 = 0{\cdot}693\,(R_1+R_2)C_1$	$t_1 \simeq 0{\cdot}7\,R_2\,C_1$
$t_2 = 0{\cdot}693\,(R_2)C_1$	$t_2 \simeq 0{\cdot}7\,R_2\,C_1$
$T = 0{\cdot}693(R_1+2R_2)C_1$	$T \simeq 1{\cdot}4\,R_2\,C_1$
$f = \dfrac{1{\cdot}44}{(R_1+2R_2)C_1}$	$f \simeq \dfrac{0{\cdot}72}{R_2\,C_1}$

Fig. 2.16. Basic circuit of 1kHz astable multivibrator with timing formulas

Fig. 2.16 shows the practical circuit of a basic 555 1kHz astable multivibrator, together with the formulas that define the timing of the circuit. Note that *trigger* pin 2 of the i.c. is shorted to the pin 6 *threshold* terminal, and that timing resistor R_2 is wired between pin 6 and *discharge* pin 7. Circuit operation is quite simple, as follows.

When power is first applied to the circuit C_1 starts to charge exponentially (in the normal monostable fashion) via the series R_1–R_2 combination, until eventually the C_1 voltage rises to 2/3 V_{CC}. At this point the basic monostable action of the i.c. terminates and *discharge* pin 7 switches to the low state. C_1 then starts to discharge exponentially into pin 7 of the i.c. via R_2, until eventually the C_1 voltage falls to 1/3 V_{CC}, and *trigger* pin 2 of the i.c. is activated. At this point a new monostable timing sequence is initiated, and C_1 starts to recharge towards 2/3 V_{CC} via R_1 and R_2. The whole sequence then repeats add infinitum, with C_1 alternately charging towards 2/3 V_{CC} via R_1–R_2 and discharging towards 1/3 V_{CC} via R_2 only.

Note in the above circuit that, if R_2 is very large relative to R_1, the operating frequency of the circuit is determined essentially by the R_2

and C_1 values, and that a virtually symmetrical output waveform is generated. The graph of Fig. 2.17 shows the approximate relationship between frequency and the C_1–R_2 values under the above condition. In practice, the R_1 and R_2 values of the circuit can be varied from

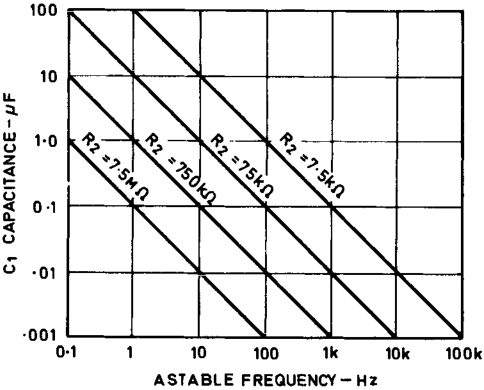


Fig. 2.17. Approximate relationship between C_1, R_2 and frequency when R_2 is large relative to R_1

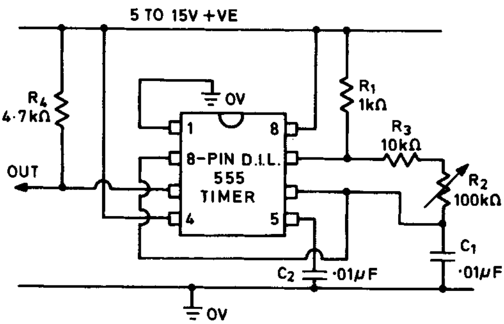


Fig. 2.18. Variable frequency square wave generator covers the range 650Hz to 7.2kHz approximately

1kΩ up to tens of megohms. Note, however, that R_1 has a significant effect on the total current consumption of the circuit, since pin 7 of the i.c. is virtually grounded during half of the timing sequence. Also note that the duty cycle or mark/space ratio of the circuit can be preset at a non-symmetrical value, if required, by suitable choice of the R_1 and R_2 values.

The basic Fig. 2.16 circuit can be usefully modified in a number of ways. Fig. 2.18, for example, shows how it can be made into a variable-frequency square wave generator by replacing R_2 with a fixed and a variable resistor in series. With the component values shown the frequency can be varied over the approximate range 650Hz – 7.2kHz via R_2.

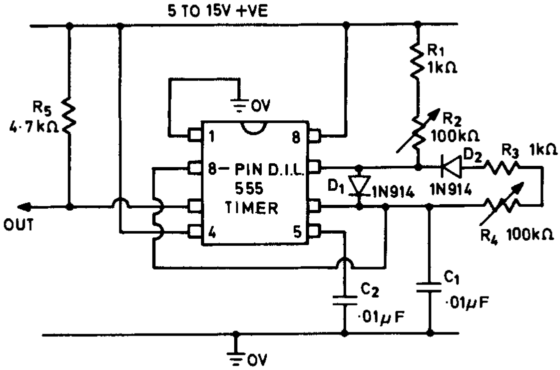


Fig. 2.19. Astable multivibrator with mark and space periods independently variable over the approximate range 7.5µs to 750µs

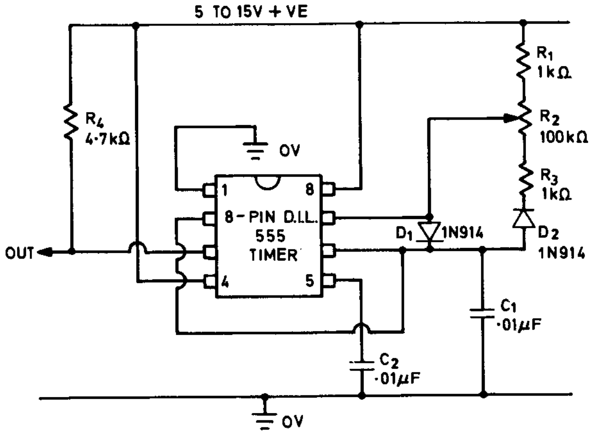


Fig. 2.20. Astable multivibrator with duty cycle variable from 1% to 99% with frequency approximately constant at 1.2kHz

Fig. 2.19 shows how the circuit can be further modified so that its *mark* and *space* periods are independently variable over the approximate range 7.5µs to 750µs. Here, timing capacitor C_1 alternately charges via R_1–R_2–D_1 and discharges via R_3–R_4–D_2.

Fig. 2.20 shows how the circuit can be additionally modified so that it acts as fixed frequency square wave generator with a mark/space ratio or duty cycle that is fully variable from 1% to 99%. Here, C_1 alternately charges via R_1 and the top half of R_2 and via D_1, and discharges via D_2–R_3 and the lower half of R_2. Note that the sum of the two timing periods is virtually constant, so the operating frequency is almost independent of the setting of R_2.

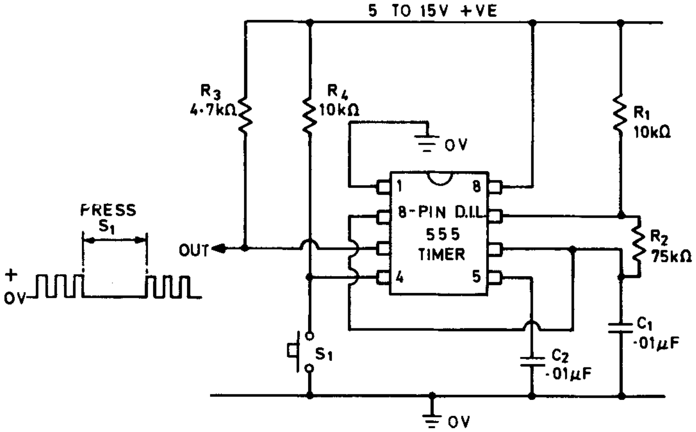


Fig. 2.21a. Gated 1kHz astable with 'press-to-turn-off' operation

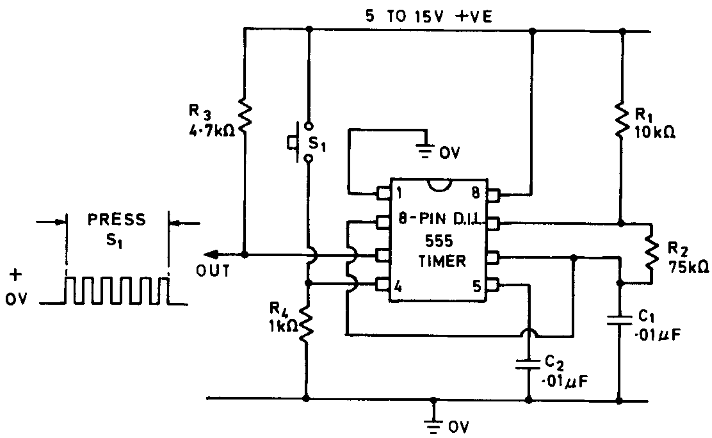


Fig. 2.21b. Gated 1kHz astable with 'press-to-turn-on' operation

The 555 astable circuit can be gated *on* or *off*, via either a switch or an electronic signal, in a variety of ways. Figs. 2.21 and 2.22 show two basic ways of gating the i.c. via a switch.

The Fig. 2.21a and 2.21b circuits are gated via the pin 4 *reset* terminal. The characteristic of this terminal is such that, if the terminal is biased significantly above a nominal value of 0.7V, the astable is enabled, but

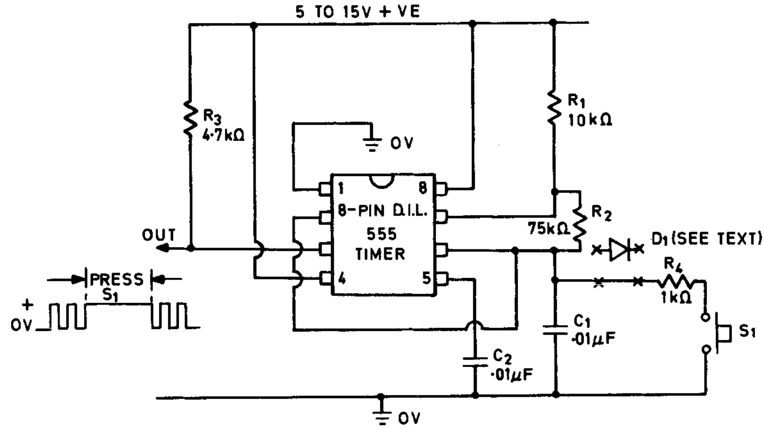


Fig. 2.22a. Alternative gated 1kHz astable with 'press-to-turn-off' operation

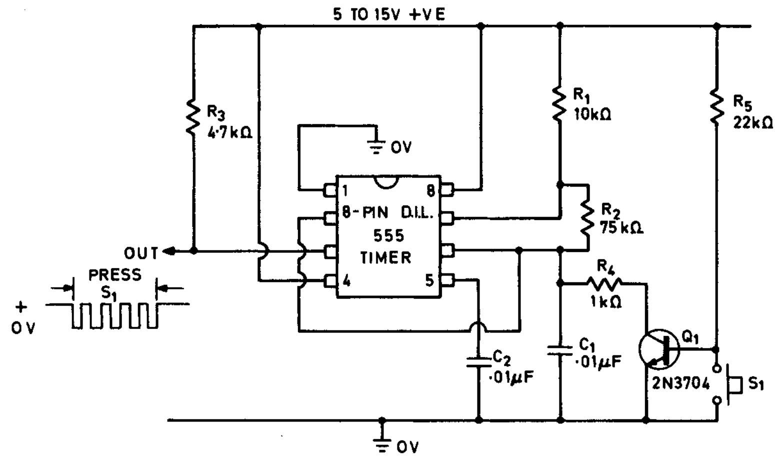


Fig. 2.22b. Alternative gated 1kHz astable with 'press-to-turn-on' operation

if the terminal is biased below 0.7V by a current greater than 0.1mA (by taking the terminal to ground via a resistance less than 7kΩ, for example) the astable is disabled and its output is grounded. Thus, the Fig. 2.21a circuit is normally on but can be turned off by closing S_1 and shorting pin 4 to ground, while the Fig. 2.21b circuit is normally gated off via R_4 but can be turned on by closing S_1 and shorting pin 4 to the positive supply rail. These circuits can alternatively be gated by applying suitable electronic signals directly to pin 4 of the i.c.

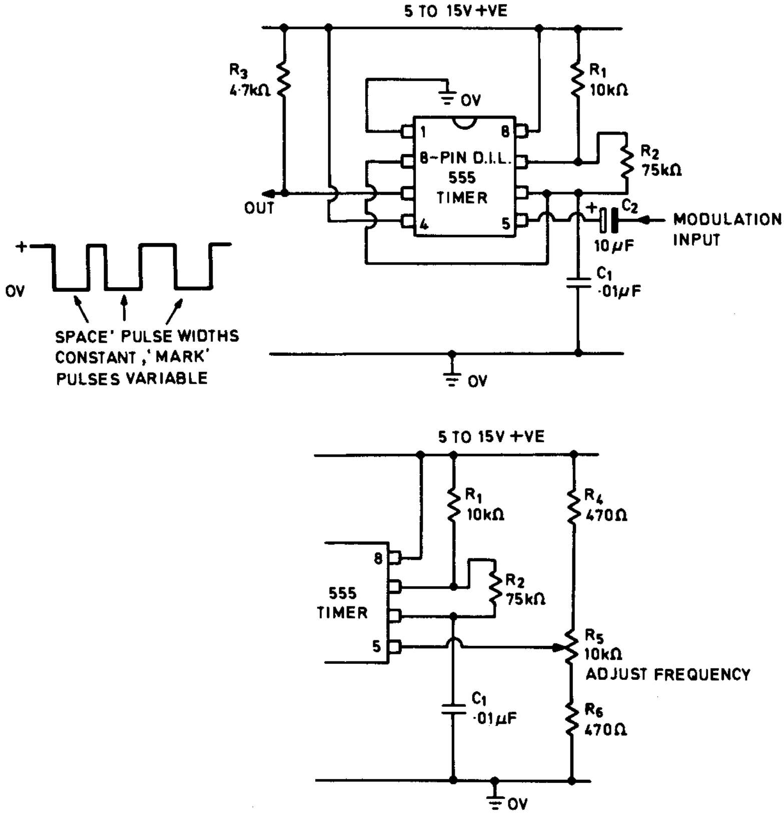


Fig. 2.23. Alternative ways of obtaining frequency or pulse-position modulation (f.m. or p.p.m.) from the 555 astable circuit

The Fig. 2.22a and 2.22b circuits are gated via the pin 2 *trigger* and pin 6 *threshold* terminals. The characteristic here is such that the circuit functions as a normal astable only as long as pin 6 is free to swing up to 2/3 V_{CC} and pin 2 is not biased below 1/3 V_{CC}. If these pins are simultaneously driven below 1/3 V_{CC} the astable action is immediately terminated and the output is driven to the high state. Thus, the Fig. 2.22a circuit is normally on but turns off when S_1 is closed. Note that an electronic signal can be used to gate the circuit by connecting a diode as indicated and eliminating S_1. In this case the circuit will gate off when the input signal voltage is reduced below $1/3\ V_{CC}$.

The Fig. 2.22b circuit is connected so that it is normally gated off by saturated transistor Q_1, but can be gated on by closing S_1 and thus turning the transistor off. This circuit can be gated electronically by eliminating R_5 and S_1 and applying a gating signal to the base of Q_1 via a 10kΩ limiting resistor. In this case the astable turns off when the input signal is high, and turns on when the input signal is reduced below 0.7V or so.

All the 555 astable circuits that we have looked at can be subjected to frequency modulation or pulse-position modulation by simply feeding a suitable modulation signal to pin 5 of the i.c. This modulation signal can take the form of an a.c. signal that is fed to pin 5 via a blocking capacitor, as in the case of Fig. 2.23a, or a d.c. signal that is fed directly to pin 5, as in the case of Fig. 2.23b. The action of the i.c. is such that the voltage on pin 5 influences the width of the 'mark' pulses in each timing cycle, but has no influence on the 'space' pulses. Thus, since the signal on pin 5 influences the position of each 'mark'

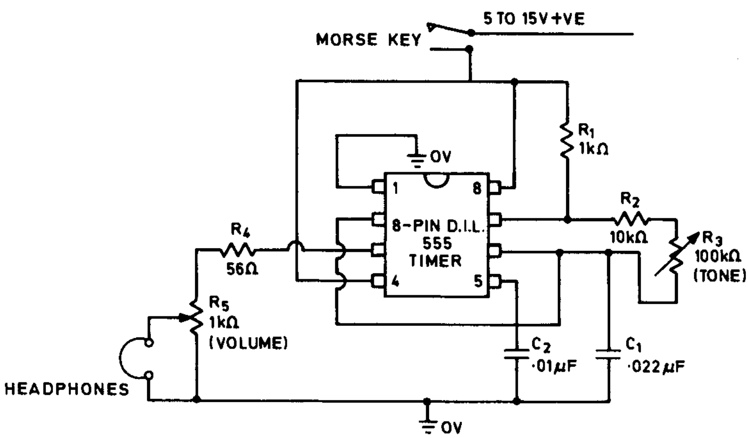


Fig. 2.24. Morse-code practice oscillator with variable tone (300Hz to 3kHz) and variable volume

pulse in each timing cycle, this terminal provides pulse-position modulation, and, since the signal influences the total period of each cycle (and thus the frequency of the output signal), the terminal also provides frequency modulation. These facilities are useful in special waveform generator applications, as is shown in the next section of this chapter.

Miscellaneous astable applications

The 555 astable multivibrator has three outstanding advantages over other types of astable circuit. First, its frequency can be varied over a wide range via a single resistive control. Second, its output has a low impedance and can source or sink currents up to 200mA. Finally, its operating frequency can readily be modulated by applying a suitable signal to pin 5 of the i.c. These features make the device exceptionally versatile, and it can be used in a vast range of practical applications of interest to both the amateur and professional user. A miscellaneous selection of such applications are shown in the present section of this chapter.

Fig. 2.24 shows how the 555 timer i.c. can be used as a morse-code practice oscillator. The circuit acts as a normal astable, with frequency

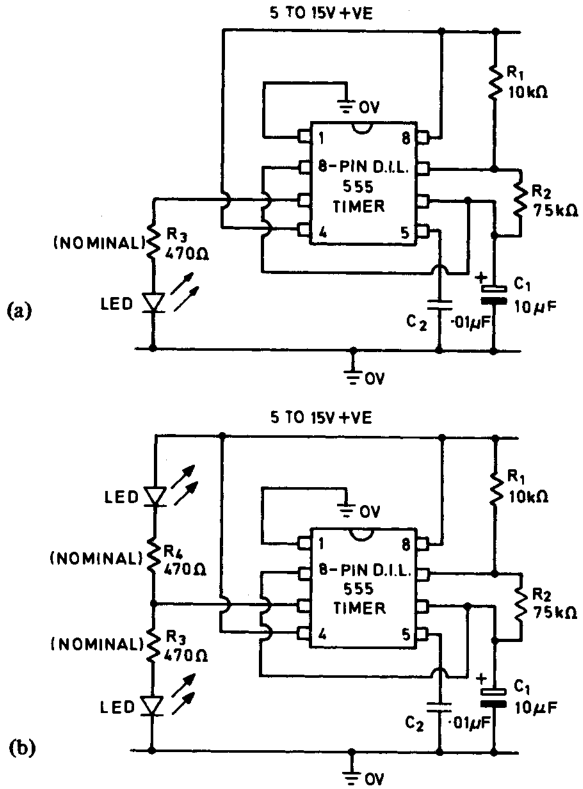


Fig. 2.25. Single output (a) and dual output (b) LED flashers give symmetrical 1Hz outputs

variable over the approximate range 300Hz – 3kHz via *tone* control R_3. The headphone volume is variable via R_5, and the headphones can have any impedance from a few ohms up to megohms. The circuit draws zero quiescent current, since the normally open morse key is used to connect the circuit to the positive supply rail, which can have any value in the range 5V to 15V.

Fig. 2.25 shows how the 555 astable circuit can be used in a couple of *LED* flasher applications. Both circuits operate at approximately 1Hz, flashing the *LEDs* on and off once every second. The Fig. 2.25a circuit has a single *LED* output. The Fig. 2.25b circuit has two *LED* outputs, and the action is such that one *LED* is on while the other is off, and vice versa. Any types of *LEDs* can be used in these circuits. Series resistors R_3 or R_4 determine the *on* current of each *LED*.

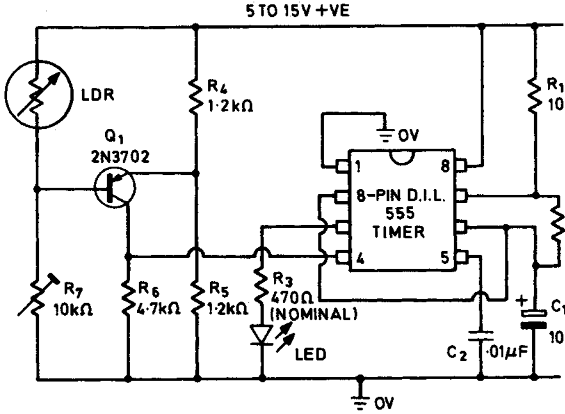


Fig. 2.26. Automatic (dark-activated) LED flasher

Fig. 2.26 shows how the Fig. 2.25a circuit can be modified to give automatic dark-activated operation. Here, R_4 and R_5 are wired as a fixed potential divider that sets 1/2 V_{CC} on the emitter of Q_1, *LDR* and R_7 are wired as a light-sensitive potential divider that applied a variable voltage to the base of Q_1, and the collector of Q_1 is taken to *reset* pin 4 of the i.c., which is normally biased to ground via R_6.

In use R_7 is adjusted so that the voltage to the base of Q_1 is greater than 1/2 V_{CC} under 'daylight' conditions, so Q_1 is cut off, but under 'dark' conditions Q_1 base is biased below 1/2 V_{CC}, so Q_1 is driven on. Thus, under daylight conditions Q_1 is cut off, so the 555 astable is disabled, with its output driven low, by 4.7kΩ resistor R_6 which is wired between pin 4 and ground. Under 'dark' conditions, on the other hand, Q_1 is biased on, so pin 4 of the i.c. is positively biased, and the astable operates normally and activates the *LED*.

The *LDR* used in the above circuit can be any cadmium-sulphide photocell that presents a resistance in the approximate range 470Ω to 10kΩ under the minimum 'dark' turn-on condition. The auto-turn-on facility of this circuit can also be used with the dual-output *LED* flasher circuit of Fig. 2.25b, if required.

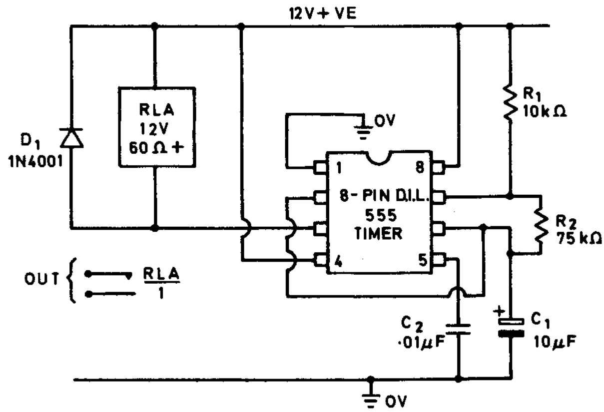


Fig. 2.27. Relay pulser can be used as emergency flasher for automobiles etc.

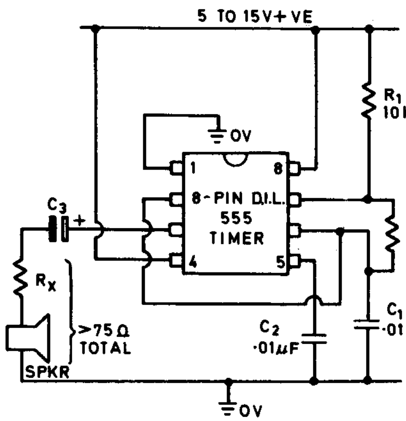


Fig. 2.28. 800Hz monotone alarm call generator

Fig. 2.27 shows how the 555 astable circuit can be used as a 12V relay pulser, which turns the relay on and off at a rate of one cycle per second. The relay can be any type with a coil resistance greater than

60Ω. This circuit can be used as the basis of an emergency light flasher in automobiles, etc.

Fig. 2.28 shows the connections for making an 800Hz monotone alarm-cell generator. This circuit can be used with any supply in the range 5 to 15V, and with any speaker impedance. Note, however, that R_X must be wired in series with speakers having impedances less than 75Ω, and must be chosen to give a total series impedance of at least 75Ω, to keep the peak speaker currents within the 200mA driving constraints of the 555 i.c. The available alarm output power of the circuit depends on the speaker impedance and supply voltage used, but may be as great as 750mW when a 75Ω speaker is used with a 15V supply.

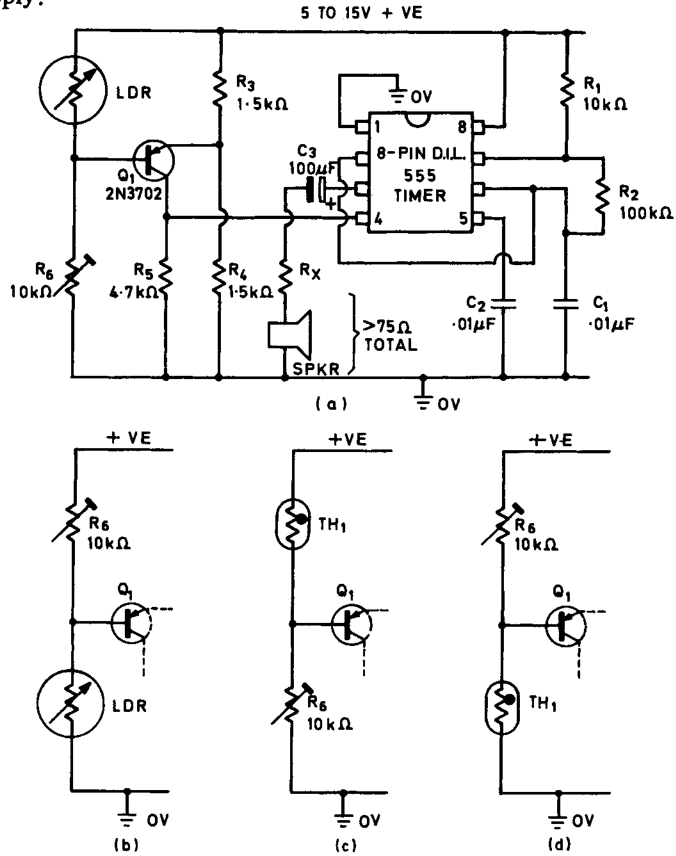


Fig. 2.29. Monotone alarm activated by dark (a), light (b), under-temperature (c) or over-temperature (d)

Fig. 2.29 shows how the above circuit can be modified so that it is activated by darkness (a), by brightness (b), by an under-temperature (c), or by an over-temperature (d). The triggering circuit, which is designed around Q_1, works in the same way as already described for the Fig. 2.26 automatic (dark-activated) *LED* flasher. The *LDR* used in the light-activated versions of this circuit can be any cadmium-sulphide photocells that present resistances in the approximate range 470Ω to 10kΩ at the desired turn-on levels. The thermistors used in the temperature-activated versions of the circuit can be any negative-temperature-coefficient types that present resistances in the same range at the required turn-on temperatures.

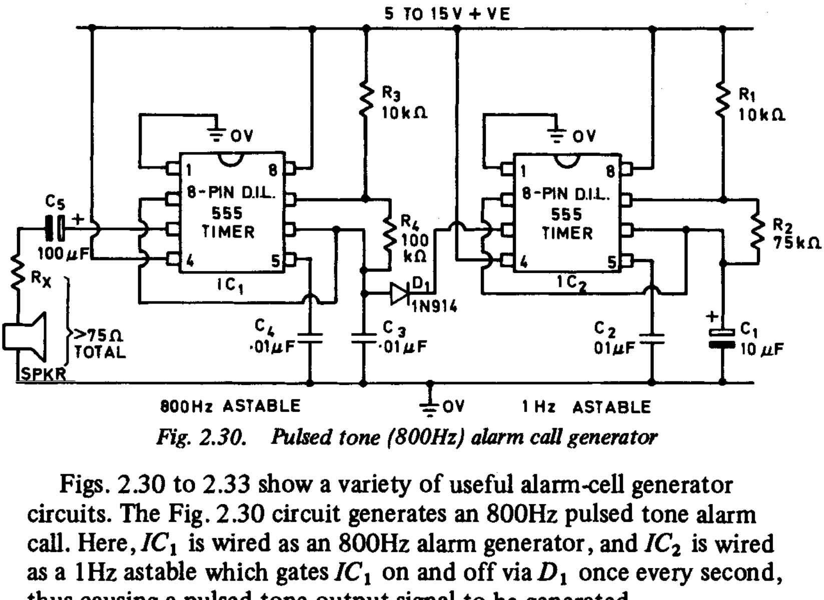


Fig. 2.30. Pulsed tone (800Hz) alarm call generator

Figs. 2.30 to 2.33 show a variety of useful alarm-cell generator circuits. The Fig. 2.30 circuit generates an 800Hz pulsed tone alarm call. Here, IC_1 is wired as an 800Hz alarm generator, and IC_2 is wired as a 1Hz astable which gates IC_1 on and off via D_1 once every second, thus causing a pulsed-tone output signal to be generated.

The Fig. 2.31 circuit generates a warble-tone alarm signal that simulates the 'dee-dah' sound of a British police siren. Here, IC_1 is again wired as an alarm generator and IC_2 is wired as a 1Hz astable multivibrator, but in this case the output of IC_2 is used to frequency modulate IC_1 via R_5. The action is such that the output frequency of IC_1 alternates symmetrically between 500Hz and 440Hz, taking one second to complete each alternating cycle.

The Fig. 2.32 circuit generates a 'wailing' alarm that simulates the sound of an American police siren. Here, IC_2 is wired as a low frequency astable that has a cycling period of about 6 seconds. The slowly varying 'ramp' waveform of C_1 of this i.c. is fed to pnp emitter follower Q_1, and is then used to frequency modulate alarm generator IC_1 via R_6. IC_1 has a natural centre frequency of about 800Hz. The circuit action is such that the alarm output signal starts at a low frequency, rises for 3 seconds to a high frequency, then falls over 3 seconds to a low frequency again, and so on add infinitum.

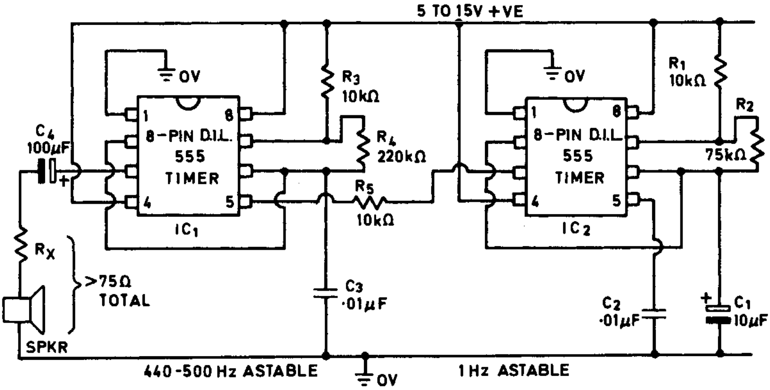


Fig. 2.31. Warble-tone alarm-call generator simulates British police siren

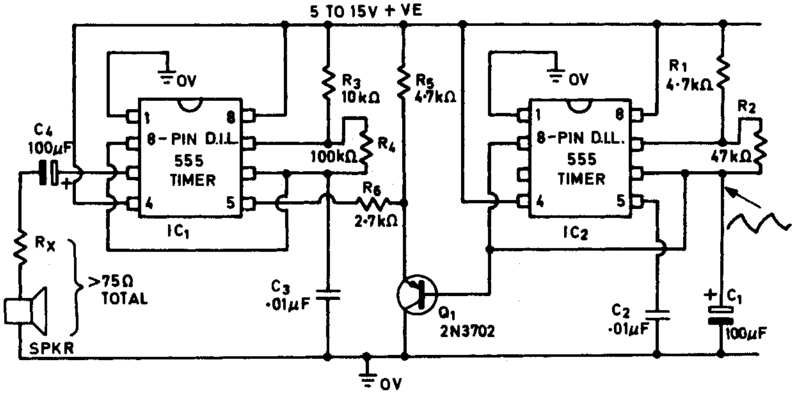


Fig. 2.32. 'Wailing' alarm simulates American police siren

Finally, to complete this quartet of alarm generator circuits, the Fig. 2.33 circuit generates a siren alarm signal that is a simulation of the 'Red Alert' alarm used in the *Star Trek* TV series. This signal starts at a low frequency, rises for about 1.15 seconds to a high frequency, ceases for about 0.35 seconds, then starts rising again from a low frequency, and so on add infinitum. The circuit action is as follows.

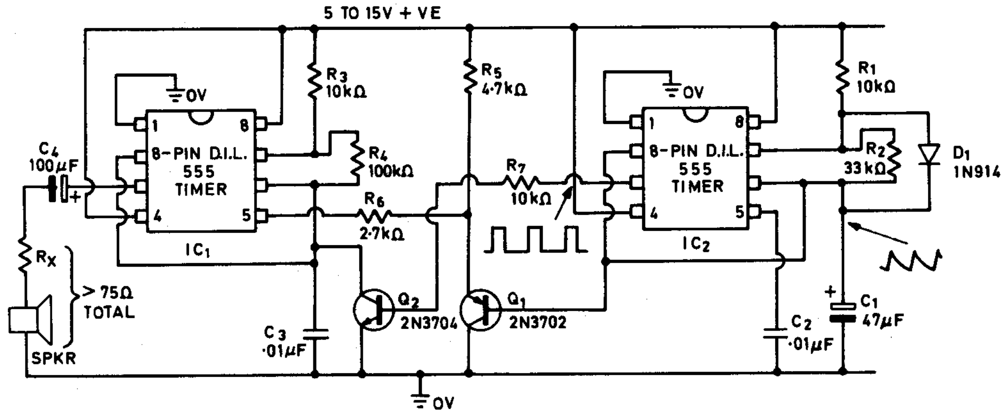


Fig. 2.33. 'Red alert' siren alarm simulates 'Star Trek' alarm signal

IC_2 is wired as a non-symmetrical astable multivibrator, in which C_1 alternately charges via R_1 and D_1, and discharges via R_2, thus giving a rapidly rising and slowly falling 'sawtooth' waveform across C_1. This waveform is fed to pnp emitter follower Q_1, and is thence used to frequency modulate pin 5 of IC_1 via R_6. Now, the frequency modulation action of pin 5 of the IC_1 astable circuit is such that a rising voltage on pin 5 causes the astable frequency to fall, and vice versa. Consequently, the sawtooth modulation signal on pin 5 causes the astable frequency to rise slowly during the falling part of the sawtooth and collapse rapidly during the rising part of the sawtooth. The rectangular pin 3 output of IC_2 is used to gate IC_1 off via npn common emitter amplifier Q_2 during the collapsing part of the signal, so only the rising parts of the alarm signal are in fact heard, as in the case of the genuine *Star Trek* 'Red Alert'.

Miscellaneous 555 applications

To complete the 555 story, this final section of this chapter shows a miscellany of 555 applications, of varying degrees of usefulness. Fig. 2.34 shows how a single type 555 i.c. can be used as the basis of an

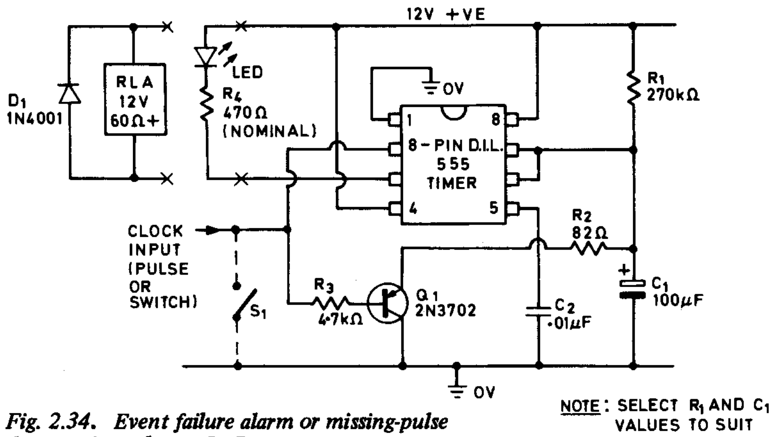


Fig. 2.34. Event failure alarm or missing-pulse detector has relay or LED output

event-failure alarm or a missing-pulse detector, which closes a relay or illuminates an *LED* if a normally recurrent event fails to take place. The circuit can be used to sound an alarm or take some other action if a monitored heart stops beating, or if a piece of machinery stops operating, or if the frequency of a waveform generator falls below a prescribed limit, etc.

The operating theory of the Fig. 2.34 circuit is fairly simple. The 555 i.c. is wired as a normal monostable pulse generator, except that transistor Q_1 is wired across timing capacitor C_1 and has its base taken to *trigger* pin 2 of the i.c. via R_3. The *trigger* pin of the i.c. is fed with a train of pulse- or switch-derived clock input signals from the monitored event, and the values of R_1 and C_1 are selected so that the monostable period of the i.c. is slightly longer than the repetition period of the clock input signal.

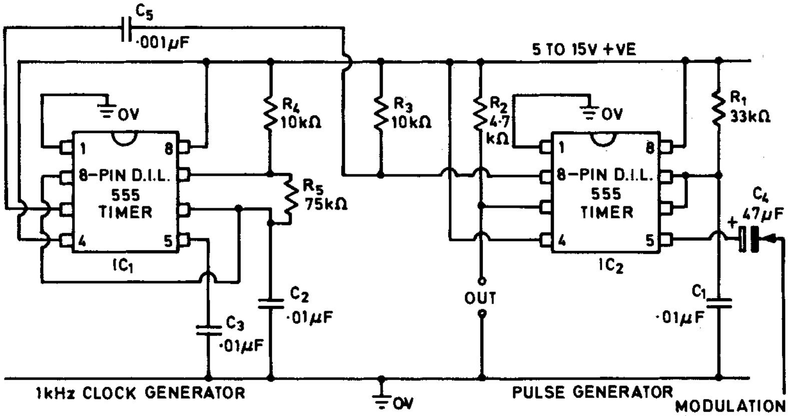


Fig. 2.35. Pulse-width modulation (p.w.m.) circuit

Thus, each time a clock pulse arrives a monostable timing period is initiated via pin 2 of the i.c., and C_1 is discharged and the pin 3 output is driven high via transistor Q_1. Before each monostable period can terminate, a new clock pulse arrives, and a new monostable period is initiated, so the pin 3 output terminal remains high so long as clock input pulses continue to arrive within the prescribed period limits. Should a clock pulse be missed, or the clock period exceed the predetermined limits, however, the monostable period will be able to terminate normally, and pin 3 of the i.c. will go low and drive the relay or *LED* on. The circuit thus functions effectively as an event-failure alarm or missing-pulse detector. With the component values shown, the monostable has a natural period of about 30 seconds. This period can be varied via R_1 and C_1 to satisfy specific requirements.

Fig. 2.35 shows how a couple of 555 i.c.s can be used to make a pulse-width modulation circuit. This circuit can be used for transmitting coded messages, or for applying variable power to a load at maximum efficiency.

Here, IC_1 is wired as a 1kHz astable multivibrator, which is used to feed a continuous train of clock pulses to the pin 2 *trigger* terminal of

IC_2, which is wired as a normal monostable multivibrator or pulse generator and has a natural monostable period of approximately 0.36ms. The external modulation signal is fed to the pin 5 *control voltage* terminal of the monostable via C_4, and determines the instantaneous widths of the generated pulses. Thus, the Fig. 2.35 circuit generates a train of pulse-width modulated (p.w.m.) pulses at a fixed repetition frequency of 1kHz.

Fig. 2.36 shows how a basic 555 monostable multivibrator can be modified so that it generates a linear ramp waveform of fixed duration each time it is triggered. This circuit can form the basis of an excellent oscilloscope timebase generator. The circuit works just like a normal monostable circuit, except that timing capacitor C_1 is charged via constant-current generator Q_1 during each timing cycle, thus causing a linear ramp voltage to be generated across C_1.

When a capacitor is charged via a constant-current generator, the voltage across the capacitor rises linearly at a predictable rate that is determined by the magnitudes of the charging current and the capacitance. The relationship can be expressed as

$$\text{Volts-per-second} = \frac{I}{C}$$

when I is expressed in amps and C is expressed in farads. Using more practical quantities, alternative expressions for the rate of voltage rise are

$$\text{Volts-per-}\mu\text{s} = \frac{\text{Amps}}{\mu\text{F}}$$

or

$$\text{Volts-per-ms} = \frac{\text{mA}}{\mu\text{F}}.$$

Note that the rate-of-rise of the voltage can be increased by increasing the charging current or decreasing the value of capacitance.

In the Fig. 2.36 circuit the charging current can be varied over the approximate range 90μA to 1mA via R_4, thus giving rates of rise on the 0.01μF capacitor of 9V-per-ms to 100V-per-ms respectively. Now, remembering that each monostable period of the 555 circuit terminates at the point when C_1 voltage reaches 2/3 V_{CC}, and assuming that a 9V supply is used (giving a 2/3 V_{CC} value of 6V), it can be seen that the monostable cycles of the Fig. 2.36 circuit have periods variable from 666μs (= 6/9-per-ms) to 60μs (= 6/100-per-ms) respectively. Periods can be increased beyond these values by increasing the C_1 value, or vice versa. Note when using this circuit that its supply rail must be stabilised if stable timing periods are to be obtained.

If the Fig. 2.36 circuit is to be used as the basis of an oscilloscope timebase, note that the input driving signal must first be converted to a good square wave, from which suitable trigger pulses can be derived via C_3 and R_5. The minimum useful ramp period that can be obtained from the circuit is about 5μs, which, when expanded to give full deflection on a ten-division oscilloscope screen, gives a maximum timebase speed of 0.5μs-per-division. Flyback beam-suppression signals can be derived from the pin 3 *output* terminal of the monostable i.c.

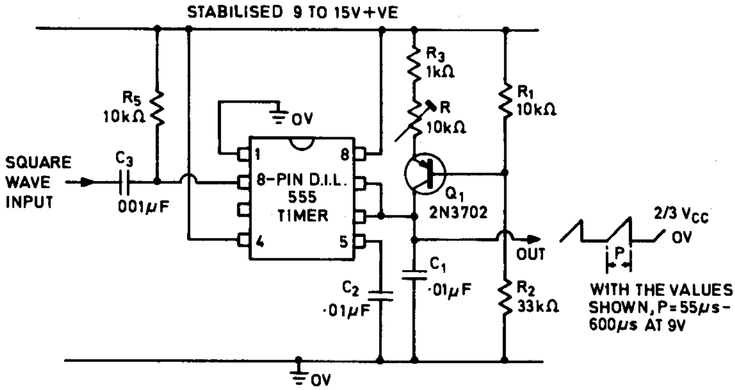


Fig. 2.36. Triggered linear-ramp generator can be used as the basis of an oscilloscope timebase

The 'timebase' circuit gives superb signal synchronisation at trigger frequencies up to about 150kHz. If the timebase is to be used with input signal frequencies greater than this, the input signals should be divided down via a single- or multi-decade digital divider. Using this technique, the timebase can be used to view input signals up to many megahertz.

Fig. 2.37 shows how a 555 timer i.c. can be connected for use as a simple but effective Schmitt trigger or sine/square converter. The circuit acts as a good converter at input frequencies up to 150kHz or more. The circuit acts by changing its output state each time the pin 2 input signal swings from above the 2/3 V_{CC} level to below the 1/3 V_{CC} level, or vice versa. Resistor R_3 is wired in series with pin 2 of the i.c. to ensure that the input signal is not adversely influenced by the transition action of the i.c.

Fig. 2.38 shows how the basic Schmitt circuit can be adapted to a dark-activated relay driving application by wiring light-dependent potential divider R_1–LDR to the pin 2 input terminal of the i.c. This circuit has an inherently high degree of input backlash, and is likely to be of value in only very specialised applications.

A far more useful relay-driving switching circuit is shown in Fig. 2.39. This circuit has negligible input backlash, and can be used as either a light- or temperature-activated switch. In light-activated applications R_1 is wired in series with a cadmium-sulphide photocell that presents a resistance in the approximate range 470Ω to 10kΩ at the required turn-

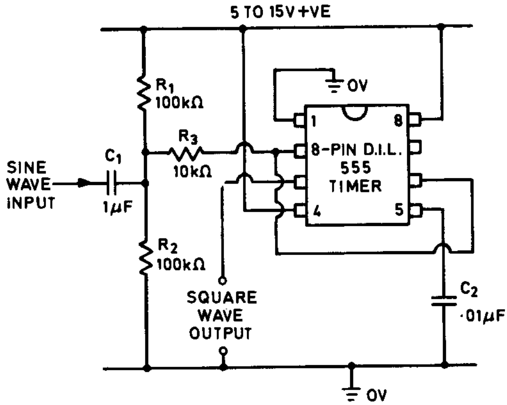


Fig. 2.37. 555 Schmitt trigger circuit acts as excellent sine/square converter up to about 150kHz

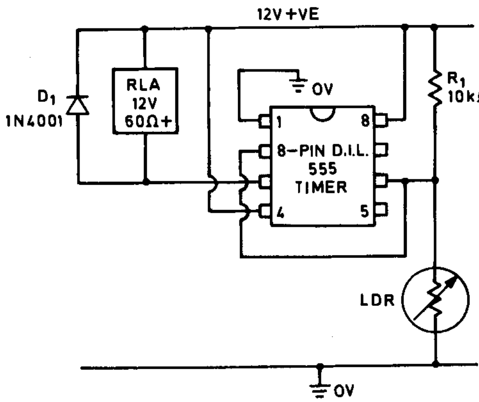


Fig. 2.38. Dark-activated relay switch has built-in backlash

on level. Dark-activated operation can be obtained by using the connections shown in Fig. 2.39a, or light-activated operation can be obtained by using the connections shown in Fig. 2.39b.

For temperature-activated operation, R_1 must be wired in series with a negative-temperature-coefficient thermistor. This thermistor must present a resistance in the range 470Ω to 10kΩ at the required turn-on level. Under-temperature operation can be obtained by using the connections shown in Fig. 2.39c, or over-temperature operation can be obtained by using the connections shown in Fig. 2.39d.

Finally, to complete this chapter, Fig. 2.40 shows how the 555 timer i.c. can be used as the basis of a simple 1kHz linear-scale analogue

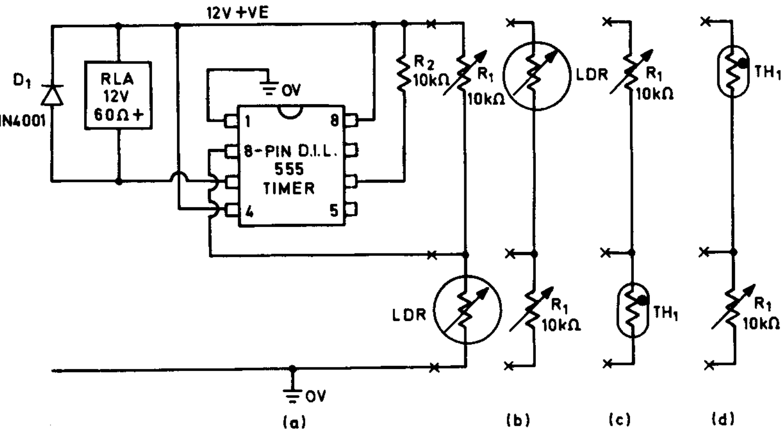


Fig. 2.39. Minimum-backlash relay switch can be activated by dark (a), light (b), under-temperature (c), or over-temperature (d)

frequency meter. The circuit needs a square-wave input driving signal with a peak-to-peak amplitude of 2 volts or greater. In this circuit the 555 i.c. is wired as a standard monostable multivibrator or pulse generator, and is powered from a regulated 6V supply. Transistor Q_1 is used to simply amplify the square wave input signals to a level suitable for triggering the monostable stage, and the output of the monostable is fed to 1mA f.s.d. meter M_1 via multiplier resistor R_5 and offset-cancelling diode D_1. This meter gives a reading that is directly proportional to the frequency of the square wave input signals, and its operating theory is as follows.

Each time the monostable multivibrator is triggered it generates a pulse of fixed duration and fixed amplitude. If we assume that each generated pulse has a peak amplitude of 10V and a period of 1ms, and that the pulse generator is triggered at an input frequency of 500Hz, it can be seen that the pulse is high (at 10V) for 500ms in each 1000ms (one second) total period, and that the *mean* value of output voltage measured over this total period is 500/1000ms x 10V = 5V, or 50% of 10V. Similarly, if the input frequency is 250Hz the pulse is high for 250ms in each 1000ms total period, so the mean output voltage equals

250/1000ms x 10V = 2.5V, or 25% of 10V. Thus, the mean value of output voltage of the pulse generator, measured over a reasonable total number of pulses, is directly proportional to the repetition frequency of the generator.

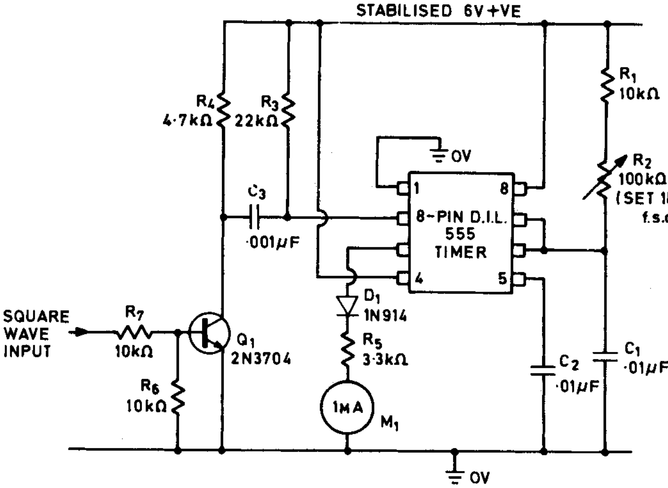


Fig. 2.40. Simple 1kHz linear scale analogue frequency meter

Normal moving coil meters are 'mean' reading instruments, and in the Fig. 2.40 circuit a 1mA f.s.d. moving coil meter is wired in series with voltage multiplier resistor R_5, which sets the meter sensitivity at about 3.4V f.s.d., and is connected so that it reads the mean output voltage of the pulse generator. This meter thus gives a reading that is directly proportional to frequency, and the circuit thus acts as a linear-scale analogue frequency meter. With the component values shown the circuit is intended to read f.s.d. at 1kHz. To set up the circuit initially, simple feed a 1kHz square wave signal to its input, and then adjust R_2 (which controls the pulse lengths) to give full-scale reading on the meter; all adjustments are then complete.

The full-scale frequency of the above circuit can be varied from about 100Hz to about 100kHz by suitable choice of C_1 value. The circuit can be used to read frequencies up to tens of MHz by feeding the input signals to the monostable circuit via a single- or multi-decade digital divider, thereby reducing the input frequencies to values that can be read by the monostable circuit. The circuit can form the basis of an excellent and inexpensive multi-range linear-scale analogue frequency meter.

CHAPTER 3

XR - 2206 WAVEFORM GENERATOR CIRCUITS

The XR-2206 integrated circuit is formally described in the manufacturer's data sheet as a monolithic function generator i.c. In less formal terms, it is a versatile and economically priced i.c. that is capable of generating high quality sine, square, triangle, ramp, and pulse waveforms, at frequencies from a fraction of a Hz to several hundred kHz, with a minimum of external circuitry. The frequency can be swept over a 2000:1 range using a single control voltage or resistance. In addition, the generator can be subjected to a.m. or f.m. control, or to phase-shift or frequency-shift keying.

The XR-2206 i.c., manufactured by Exar Integrated System, Inc., is available from many distributors. The i.c. is housed in a standard 16-pin d.i.l. package, and can be powered from either a single supply in the range 10V to 26V, or a split supply in the range ±5V to ±13V. When used in the sine wave generator mode, the t.h.d. of the signal is typically 2.5% without adjustment, but can be reduced to about 0.5% via external trimmer controls. The sine wave output signal has a typical maximum amplitude of 2V r.m.s. and an output impedance of 600Ω. The frequency stability of the i.c. is excellent, being of the order of 20ppm/°C for thermal changes and 0.01%/V for supply voltage changes.

The XR-2206 integrated circuit has many useful applications, and can readily be used as a simple waveform generator or as a complex function generator with a variety of modulation facilities.

How it works

The XR-2206 i.c. is housed in a 16-pin d.i.l. package. Fig. 3.1 shows the functional block diagram and pin connections of the device. The heart of the unit is a voltage-controlled oscillator (v.c.o.), which is driven via a pair of current switches. The v.c.o.s main timing capacitor is wired between pins 5 and 6, and can have any value in the range 1000pF to 100μF. The v.c.o.s main timing resistor (which controls the device's

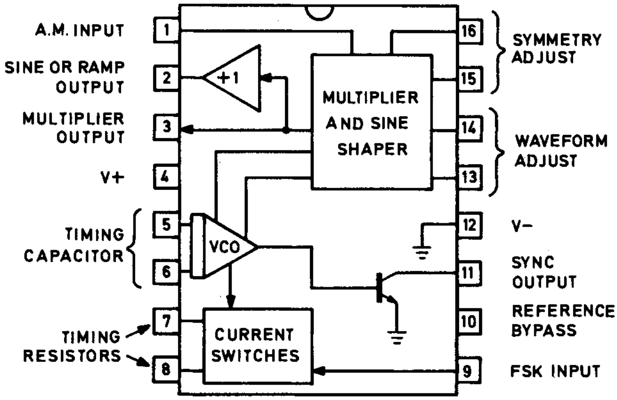


Fig. 3.1. Functional block diagram and pin connections (top view) of the XR-2206 function generator i.c.

timing currents) is wired between the negative supply rail of the circuit and pin 7 or pin 8 of the i.c., and can have any value in the range 1kΩ to 2MΩ.

The frequency of oscillation, f_o, is determined by the external timing capacitor across pins 5 and 6, and by the timing resistor R connected to either pin 7 or pin 8. The frequency is given as

$$f_o = 1/RC \text{ Hz}$$

and can be varied via either R or C. For optimum thermal stability and minimum sine wave distortion, R should have a value in the range 4kΩ to 200kΩ.

Either a pin 7 or pin 8 timing resistor can be selected by applying a suitable voltage or signal to the pin 9 *FSK Input* terminal of the i.c. If pin 9 is open-circuited or connected to a bias voltage greater than 2V, the pin 7 resistor is selected. Conversely, if pin 9 is biased below 1V, the pin 8 timing resistor is selected. This *FSK* facility enables the output signal to be switched alternately between two independently adjustable frequencies, to produce, for example, a warble-tone signal.

The v.c.o. section of the i.c. produces two basic waveforms simultaneously. One of these is a linear ramp, which is fed to an internal multiplier and sine shaper block, and the other is a rectangular waveform, which can be passed on to pin 11 via a built-in buffer transistor. In very simple terms, the action of the v.c.o. is such that the timing capacitor first charges linearly via the timing resistor to produce a rising ramp at one output and a 'high' rectangle voltage at the other, until a certain 'firing' voltage is reached, at which point the rectangle output switches sharply to the 'low' state and the timing capacitor starts charging in the reverse direction via the timing resistor to produce a falling output ramp. The ramp continues to fall until a second firing voltage is reached, at which point the rectangular output switches sharply back to its original 'high' state, and the whole timing process then repeats add infinitum.

Thus, the v.c.o. produces symmetrical triangular and square waveforms if the same timing resistor is used to control both charging cycles of the timing capacitor. Alternatively, if the pin 11 rectangular output waveform is shorted to the pin 9 *FSK* terminal of the i.c. the v.c.o. automatically switches between the pin 7 and pin 8 timing resistors on alternate half cycles, thus enabling the i.c. to simultaneously produce non-symmetrical linear ramp (or sawtooth) and non-symmetrical square (or pulse) output waveforms.

The v.c.o. section of the i.c. is actually a current-controlled multivibrator, in which the timing current is controlled by the resistors connected to pins 7 or 8, or by external voltages or signals that are connected to these pins via suitable current-limiting resistors. This facility makes it possible to externally frequency modulate or frequency sweep the generated signals.

The ramp output waveform of the v.c.o. section of the XR-2206 i.c. is fed into the multiplier and sine shaper block of the device. This block acts like a gain-controlled differential amplifier which effectively gives a high impedance output at pin 3 and a 600Ω buffered output at pin 2. With pins 13 and 14 open, a ramp output waveform is available at pins 2 and 3 of the i.c. With a resistance of a few hundred ohms between pins 13 and 14 the block exponentially cuts off the peaks of the ramp input signals from the v.c.o., thus producing a sine wave output from pins 2 and 3. With suitable adjustment, the sine waveform distortion can typically be reduced to a mere 0.5%.

The gain and output phase of the multiplier can be varied by applying a bias or signal voltage to pin 1 of the i.c. The output is linearly controlled by variations in the pin 1 voltage around the half-supply potential level. The output is zero when the pin 1 voltage is at half-supply value, and rises as the voltage increases. When the voltage is reduced below the half-supply value, the output signal level again increases, but its phase is reversed. This characteristic can be used to amplitude modulate or phase shift key the outputs of the waveform generator at pins 2 and 3.

The effectively high output impedance of pin 3 of the XR-2206 i.c. is connected to the input of a built-in unity gain amplifier stage, which produces a buffered 600Ω output at pin 2. Consequently, the input signal to the buffer amplifier (and hence the output at pin 2) can be effectively varied by potential divider action by wiring a variable resistor or impedance between pin 3 and an effective ground point. This facility can be utilised to provide simple gain control of the output, or can be used to facilitate gate keying or pulsing of the pin 2 output signal.

A final point to note about the XR-2206 i.c. is that the d.c. level at output pin 2 is approximately the same as the d.c. bias voltage at pin 3. Thus, d.c. level shifting can be applied to the pin 2 output by applying a suitable bias to pin 3. In most applications, pin 3 is biased half way between the positive and negative supply rail voltages. In split-supply circuits this means that the output signal swings about the zero volts (common) line.

Basic waveform generator circuits

The XR-2206 is a very versatile i.c., and is quite simple to use. Fig. 3.2 shows the connections for using it as a simple wide-range sine-wave

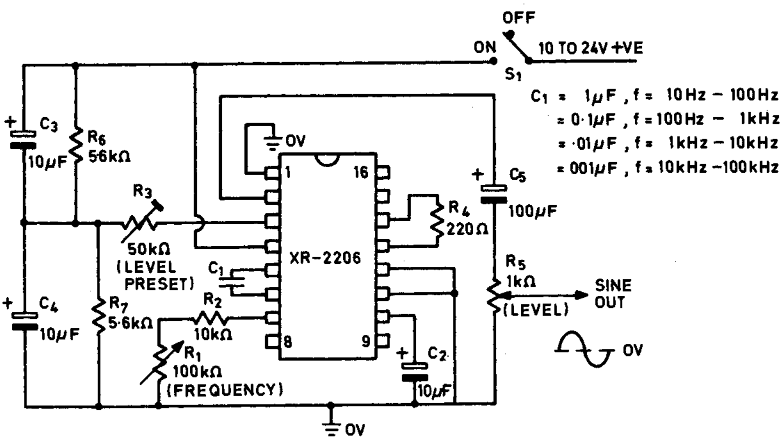


Fig. 3.2. Simple single-supply sine-wave generator

generator that is powered from a single supply source. Here, the main timing resistance comprises R_1 and R_2 in series, and enables the frequency of oscillation to be varied over more than a decade range with any given value of timing capacitor. When C_1 has a value of 1μF, the frequency can be varied from 10Hz to 100Hz via R_1, and when C_1 has a value of 0.001μF the frequency can be varied from 10kHz to 100kHz. Note that the timing resistance is connected to pin 7 of the i.c. This timing pin is automatically selected, since the pin 9 FSK terminal of the i.c. is unbiased. The circuit generates a sine wave output at pin 2, since a 220Ω resistor is wired between pins 13 and 14 of the i.c. Typically, the sine wave distortion is less than 2.5% with this simple connection.

The voltage to pin 3 of the i.c. is biased at half-supply volts by potential divider R_6–R_7, which is shunted to a low impedance by C_3–C_4, so the pin 2 sine wave signal is biased at approximately half-supply volts. This output signal is decoupled to d.c. via C_5 and made fully variable in amplitude via R_5, which provides the final output signal. The maximum amplitude of the output sine wave can be preset via level control R_3. To set up R_3, first disconnect R_4 from pin 13, so that a triangular output waveform is obtained, then reduce the R_3 value so that all clipping is removed from the triangular waveform. Note the R_3 setting. Now reconnect R_4 to pin 13, and check that a good sine wave output is obtained. The maximum sine wave amplitude can be reduced by moving R_3 below the noted setting, but must not be increased by moving R_3 above this setting. Sine wave distortion may be reduced below the typical 2.5% value by carefully adjusting the value of R_4, if desired.

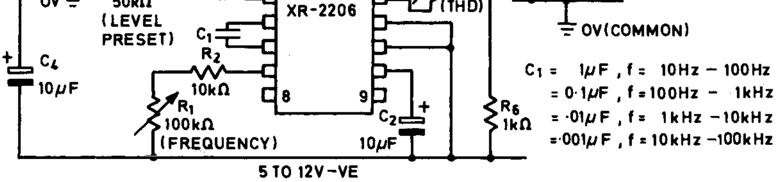


Fig. 3.3. High-performance split-supply sine-wave generator

The Fig. 3.2 circuit can be used with any supply voltage in the range 10V to 24V. The circuit can be modified for split-supply operation by replacing all ground connections with negative-rail ones, and by taking *level* control R_3 directly to the common or ground line, as shown in the high-performance split-supply sine wave generator circuit of Fig. 3.3.

Note that the R_6–R_7 potential divider is eliminated, and that decoupling capacitor C_5 is no longer required, since the sine wave output signal is automatically centred on the zero volts output line.

The Fig. 3.3 circuit also shows how the total harmonic distortion (t.h.d.) of the sine wave signal can be reduced to a typical value of 0.5% with the aid of R_4 and R_5. Here, R_4 is a preset t.h.d. adjuster, and R_5 is a preset symmetry adjust control. These controls must be adjusted alternately to give the best sine wave output waveform, after first setting R_3 to give a non-clipped triangle waveform as described above.

In the absence of a distortion factor meter, the simple twin-T 1kHz filter of Fig. 3.4 can be used in conjunction with an oscilloscope or millivoltmeter to set the sine wave generator for minimum distortion

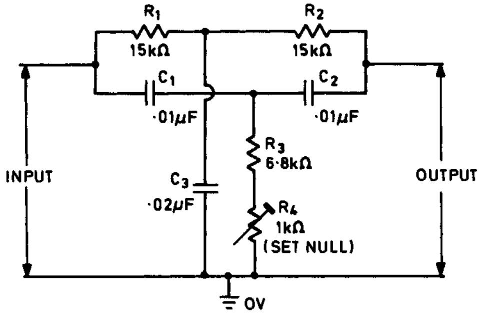


Fig. 3.4. Simple 1kHz twin-T filter aids setting up for minimum t.h.d.

at 1kHz. The procedure for using the filter is to apply the sine wave output of the generator to the input of the filter at 1V r.m.s. at approximately 1kHz, and take the output of the filter to the input of an indicator such as an oscilloscope or millivoltmeter. Next, adjust the input frequency of the generator and R_4 of the filter to give minimum output indication, and finally, adjust the R_4 and R_5 'distortion' controls of the generator to reduce the output indication of the filter to the minimum possible value. At final balance, the output of the filter corresponds to approximately 0.1% t.h.d. per mV r.m.s. of indicated reading, i.e., if the indicator shows a reading of 5mV r.m.s., the t.h.d. of the generator approximates 0.5%.

When using the low-distortion sine wave facility, it may be noted that the signal appearing at pin 3 of the i.c. is similar to that of output pin 2, but has lower distortion and a higher output impedance than that of pin 2. Also, the pin 3 signal is closely centred on the common or ground line, while the pin 1 signal is offset by a few hundred millivolts.

The pin 3 terminal can also supply a greater undistorted output signal voltage than the pin 2 terminal.

If required, d.c. offset can be applied to the pin 2 and pin 3 output signals of the Fig. 3.3 circuit by applying d.c. bias to the 'ground' side of R_3 and decoupling the bias to a.c. signals, as shown in the add-on circuit of Fig. 3.5.

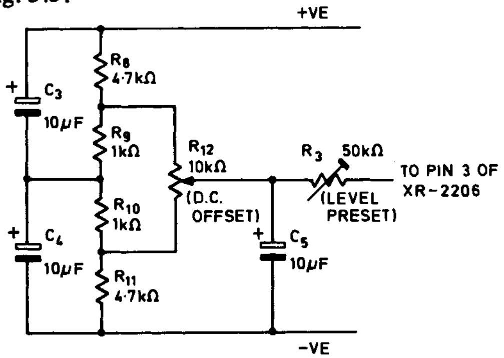


Fig. 3.5. Add-on modification for applying d.c. offset to pins 2 and 3 of the Fig. 3.3 circuit

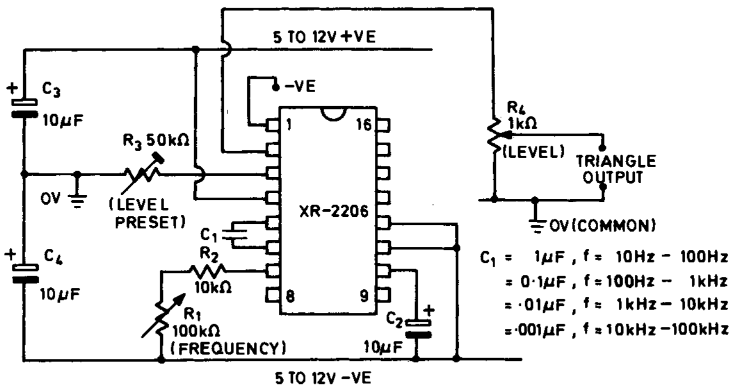


Fig. 3.6. Variable frequency triangle wave generator

The XR-2206 i.c. can be made to generate linear triangle waveforms by using the basic circuits of Figs. 3.2 and 3.3 without their sine wave shaping resistors. Fig. 3.6 shows the practical circuit of a variable-frequency split-supply triangular waveform generator. When used as a triangular waveform generator with a 9V-0-9V supply, the i.c. can typically produce signals with unloaded amplitudes of about 12V peak-to-peak before clipping occurs.

The XR-2206 i.c. can be made to produce fixed-amplitude square wave signals at pin 11, either independently or simultaneously with sine or triangular waveforms, by simply wiring a load resistor between pin 11 and the positive supply line. Fig. 3.7 shows a simple variable-frequency split-supply circuit that produces square waves only. Pin 11

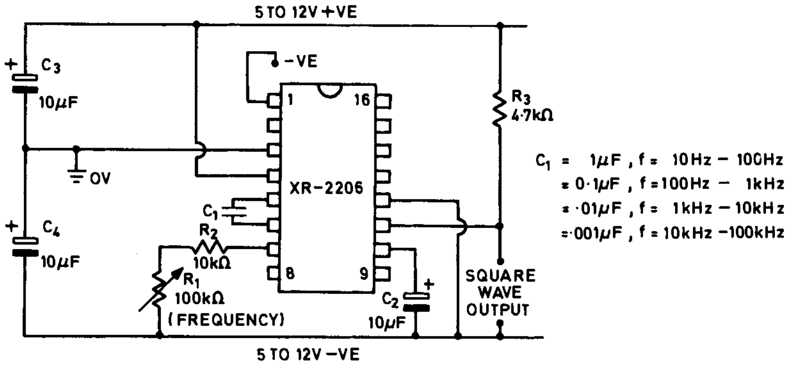


Fig. 3.7. Simple variable frequency square-wave generator

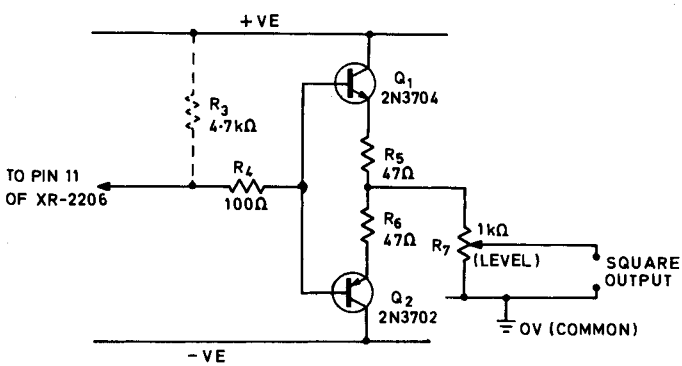


Fig. 3.8. Add-on circuit gives low-impedance variable-amplitude access to the square-wave output of the XR-2206

is referred to in the data sheets as a 'sync output' terminal, since the signal appearing at this point is not suitable for directly driving low-impedance loads but is intended only for driving high-impedance loads such as oscilloscope input or synchronisation terminals, etc. The rise and fall times of the square wave output signals are typically 250ns and 50ns respectively when pin 11 is loaded by 10pF.

Fig. 3.8 shows a simple add-on buffer stage that can be used to give low-impedance variable-amplitude access to the pin 11 square wave signal of the XR-2206. The circuit is a simple un-biased complementary emitter follower, which is driven directly from the pin 11 square wave output of the i.c., and has short-circuit output protection via the 47Ω resistors in series with the transistor emitters. The output

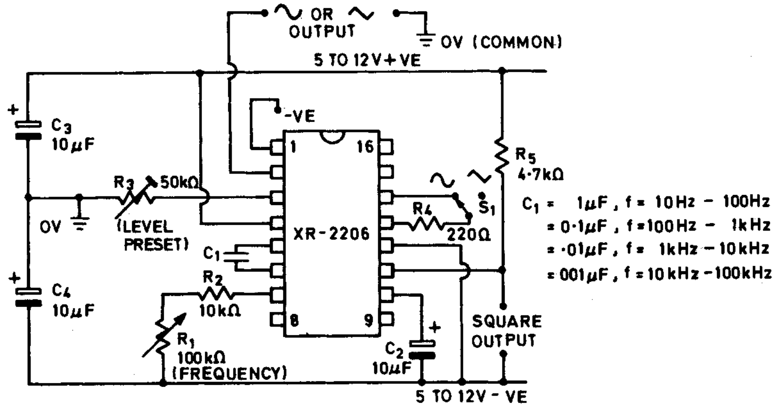
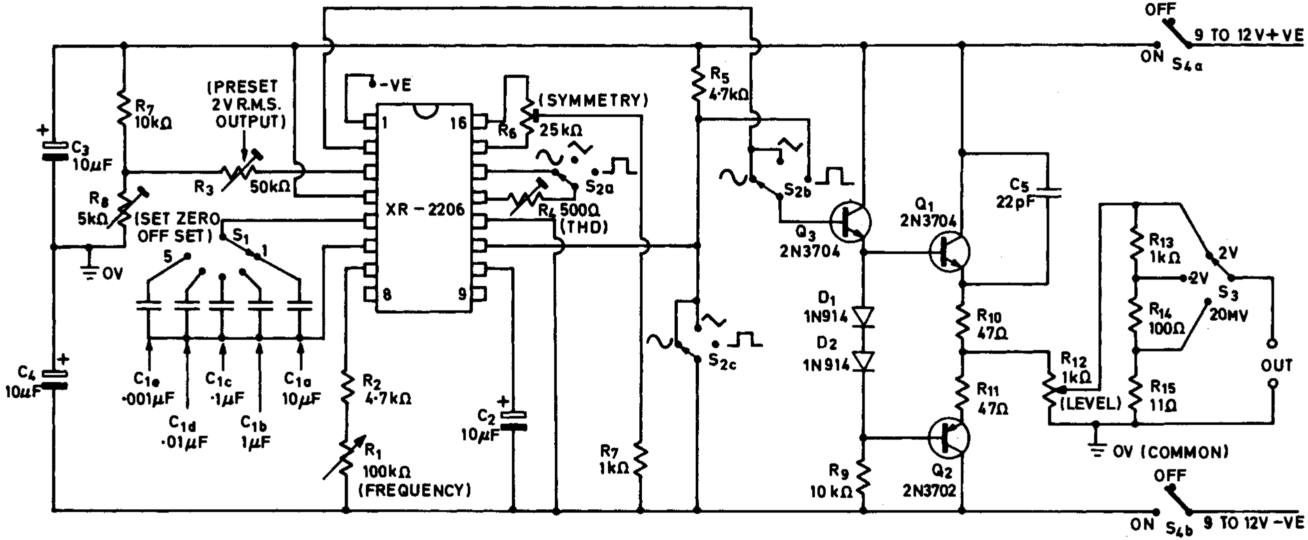


Fig. 3.9. Simple fixed-amplitude sine/triangle/square generator

level of the Fig. 3.8 circuit is fully variable from maximum to zero via the 1kΩ output potentiometer, is referenced to the zero volts or ground line, and can be used to drive high or low-impedance external loads.

The circuits of Figs. 3.2 to 3.8 can be combined in a variety of ways to make different types of waveform generators. Fig. 3.9, for example, shows how some of the circuits can be put together to make a simple fixed-amplitude *sine/triangle/square* generator. Here, the square wave output is taken directly from pin 11 of the i.c., and is produced simultaneously with the sine or triangular waveforms of pin 2, which are selected via switch S_1.

Alternatively, Fig. 3.10 shows how some of the circuits can be put together to help make a low-cost high-performance sine/triangle/square waveform generator that covers the frequency range 1Hz to 200kHz in five switched ranges. Frequencies are selected via range switch S_1 and 'fine frequency' control R_1. Each range of S_1 covers a full decade of frequency plus 100% over-range at its upper frequency. The circuit incorporates t.h.d. adjustment of the sine wave, and give a typical distortion factor of 0.5%.



S_1 ranges
Position 1: 1Hz-20Hz
2: 10Hz-200Hz
3: 100Hz-2kHz
4: 1kHz-20kHz
5: 10kHz-200kHz

Note: All capacitors are non-polarised types.

Waveform stability (typ.)
0.002 % per °C
0.01%/V supply sensitivity

Output waveforms
Sine: distortion (typ.) = 0.5%
Triangle: linearity (typ.) = 1%
Square: rise time (typ.) = 200ns
fall time (typ.) = 100ns

Total current consumption (typ.) = 20mA

S_1 = Single-pole 5-way switch
S_2 = Three-pole 3-way switch
S_3 = Single-pole 3-way switch
S_4 = Two-pole 2-way switch

Max. output levels
(with 9-0-9 V supply)
Sine = 5.6 V pk-pk
Triangle = 12 V pk-pk
Square = 16 V pk-pk

Fig. 3.10. Circuit and data of low cost high-performance 1Hz to 200kHz sine/triangle/square generator

The sine/triangle output of the i.c. is taken from pin 2, and all outputs are taken to a simple variable attenuator network via the Q_1–Q_2–Q_3 compound emitter follower stage. R_8 enables the sine/triangle output to be centred on precisely zero volts, and R_3 enables the maximum sine wave output to be set at 2V r.m.s. The performance of the circuit is summarised in the diagram. The procedure for initially setting up the circuit when it is first built is as follows.

First, set the attenuator controls to give maximum output, set the circuit to sine wave mode at about 1kHz, and then adjust R_8 to give zero offset of the output signal. This can be achieved by connecting a 0 – 2.5V d.c. meter to the output of the circuit, and then adjusting R_8 for zero reading on the meter. Next, connect a 0 – 2.5V a.c. meter to the output of the circuit, and adjust R_3 to give a sine wave output of approximately 2V r.m.s. Finally, adjust R_4 and R_6 to give minimum distortion of the sine wave, as already described, and then recheck the d.c. offset and output amplitude. The setting up procedure is then complete, and the circuit is ready for use.

Pulse and ramp generation

Fig. 3.11 shows the practical connections for using the XR-2206 as a variable-slope ramp generator circuit. Here, square wave output pin 11 is shorted to *FSK input* pin 9 of the i.c. The circuit action is such that,

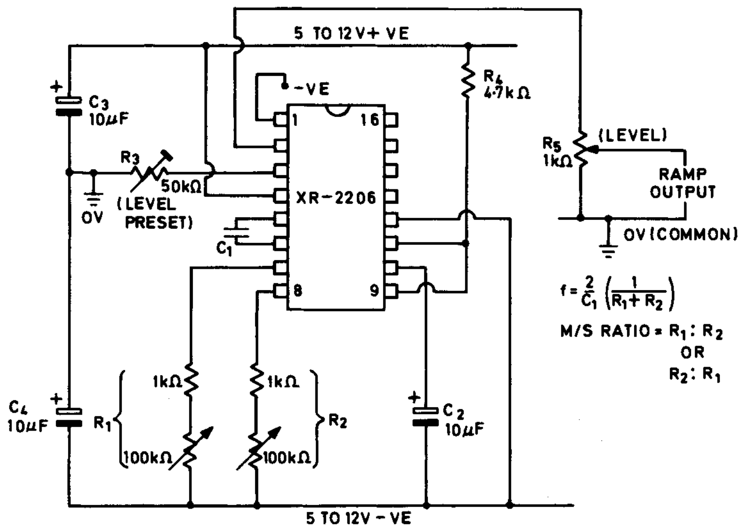


Fig. 3.11. Variable-slope ramp generator

when the pin 11 output is high, timing capacitor C_1 charges via R_1 until a point is reached at which pins 11 and 9 switch abruptly to the low state, at which point the timing capacitor recharges in the reverse direction via R_2 until pins 11 and 9 go high again, and the process then repeats add infinitum. The circuit thus automatically switches between alternate timing resistors on alternate half cycles, and produces a linear ramp output waveform at pin 2 of the i.c. The 'rise' and 'fall' periods of this waveform are independently controlled via R_1 and R_2. The operating frequency of the circuit is given as

$$f = \frac{2}{C_1}\left\{\frac{1}{R_1 + R_2}\right\}.$$

A pulse or variable mark-space-ratio 'square' wave is developed at pin 11 of the Fig. 3.11 circuit simultaneously with the ramp waveform, but

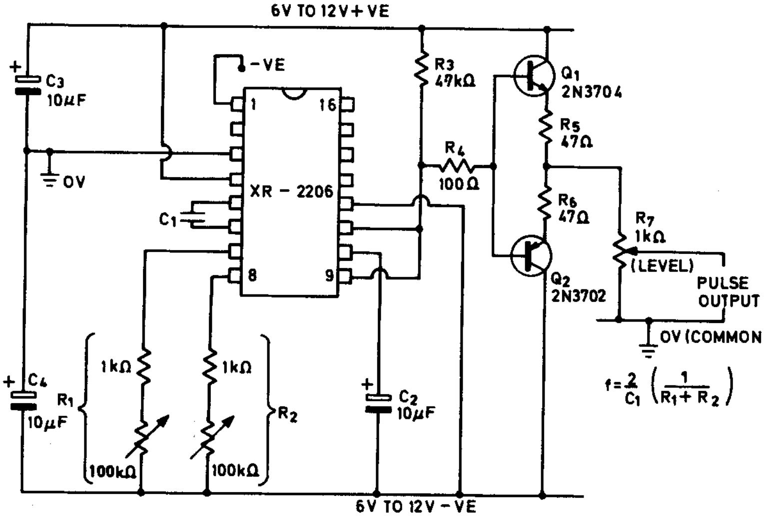


Fig. 3.12. Pulse or variable mark/space-ratio generator, with output buffer

is of fixed amplitude and is not suitable for directly driving low-impedance external loads. If required, the XR-2206 can be used specifically as a pulse or variable mark-space-ratio generator by using the connections shown in Fig. 3.12. Here, the level preset control is eliminated from the pin 3 terminal of the i.c., and the pin 11 pulse signal is made available to external loads via the Q_1–Q_2 emitter follower stage and via level control R_7.

Alternatively, the Fig. 3.11 and 3.12 circuits can be combined to form a useful variable pulse and ramp generator by using the connections shown in Fig. 3.13. Here, the desired waveform can be selected via S_1, the waveform shapes are variable via R_1 and R_2, and the waveform amplitudes are variable via R_8.

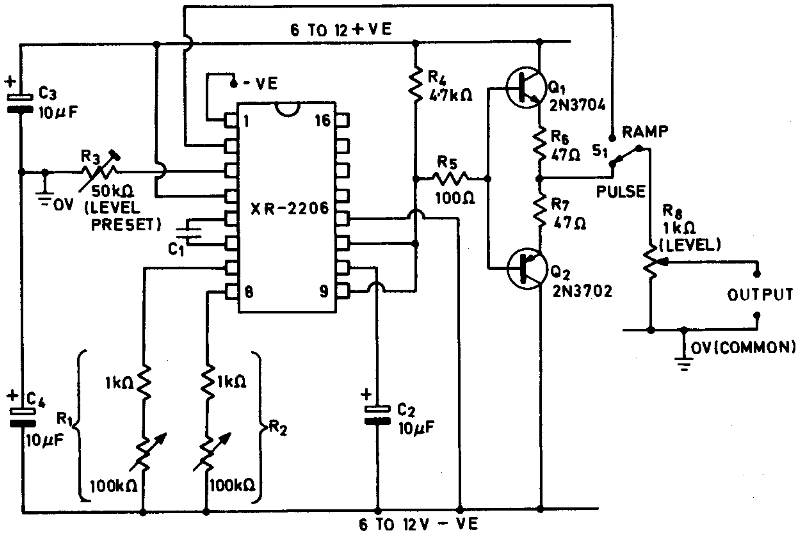


Fig. 3.13. Variable pulse and ramp generator circuit

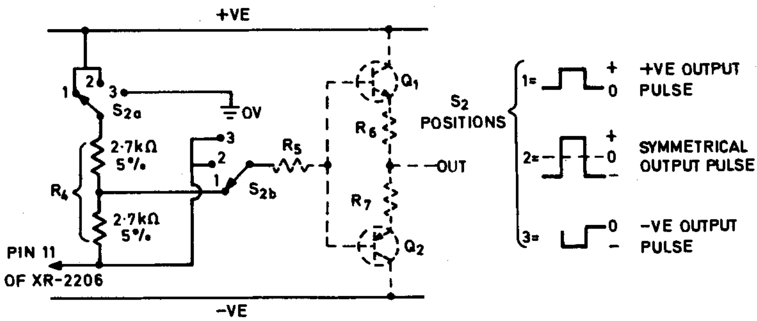


Fig. 3.14. Alternative output stage, gives +ve, −ve or symmetrical pulses

The versatility of the Fig. 3.13 circuit can be increased, if desired, by replacing the existing pulse output stage with that shown in Fig. 3.14. This circuit enables either +ve, −ve, or symmetrical output pulses to be selected via S_2. Here, load resistor R_4 is replaced by a pair of 2.7kΩ 5%

resistors. With two-pole switch S_2 in position 1 the pulse ouput of the circuit is effectively taken from the junction of these two resistors, so the output switches between the fully positive and half-supply or 'ground' volt levels, and the circuit thus gives positive output pulses. In position 2 of S_2 the output is effectively taken from pin 11 of the i.c., so symmetrical output pulses are available. In position 3 of S_2 the output is again taken from pin 11 of the i.c., but the top end of R_4 is connected to the zero volts line, so the output switches between the zero and negative voltage rails, and negative output pulses are available from the circuit.

FSK generation

When the pin 9 *FSK input* terminal of the XR-2206 i.c. is open circuit or externally biased above 2V with respect to the negative supply rail, the pin 7 timing resistor is automatically selected and the circuit operates at a frequency determined by R_1 and C_1. When pin 9 is shorted to the

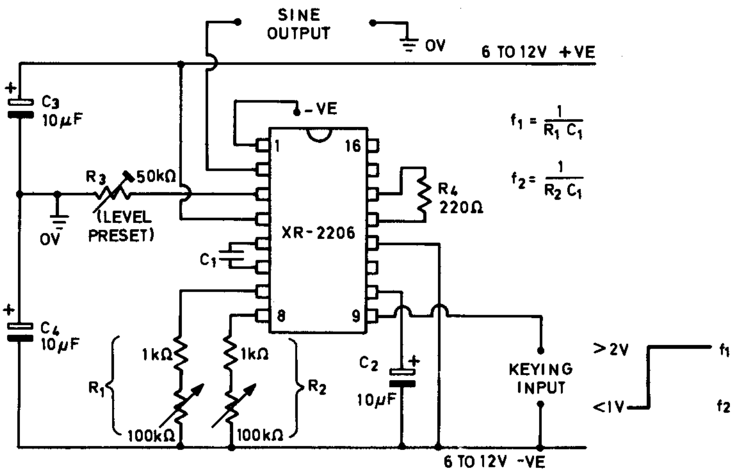


Fig. 3.15. Basic split-supply FSK sine-wave or 'warble-tone' generator

negative supply rail or is biased below 1V with reference to the negative rail, the pin 8 timing resistor is automatically selected and the circuit operates at a frequency determined by R_2 and C_1. The XR-2206 i.c. can thus be frequency-shift keyed (FSK) by simply applying a suitable keying or pulsing signal between pin 9 and the negative supply rail. Fig. 3.15 shows the practical connections for making a simple sine wave generating FSK or 'warble-tone' generator.

If required, the *FSK input* keying signal can be referenced to the ground or zero-volts line by using the three-transistor add-on circuit shown in Fig. 3.16. Here, with the input signal in either the 'low' or zero state, all three transistors are cut off, so pin 9 of the i.c. is effectively open circuit, and the operating frequency of the i.c. is controlled

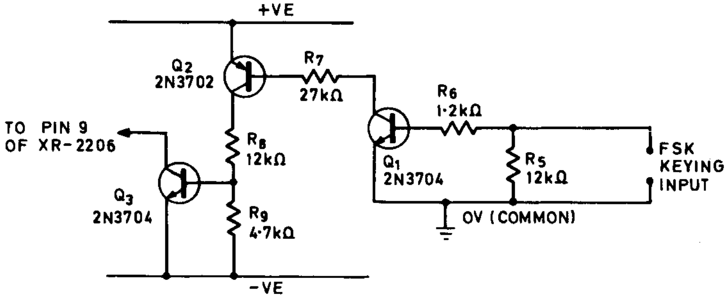


Fig. 3.16. Add-on circuit enables FSK input keying signal to be referenced to the common or ground line

by timing resistor R_1. When the *FSK input* keying signal is high, Q_1 is driven on and applies bias to Q_2, Q_2 is driven on and applies bias to Q_3, and Q_3 is driven to saturation and effectively shorts pin 9 of the i.c. to the negative supply rail, so the frequency of the i.c. is controlled by R_2. The *FSK* facility controls both the pin 2 and pin 11 signal output terminals of the circuit.

A.M. generation

The output amplitude of the pin 2 signal of the XR-2206 can be modulated by applying a d.c. bias and a modulating signal to pin 1 of the i.c., as shown in Fig. 3.17. The internal impedance at pin 1 is approximately 200kΩ, so this pin should be strapped to the negative supply rail to prevent unwanted pick-up when the a.m. facility is not in use.

The output amplitude of the pin 2 signal varies linearly with the applied voltage at pin 1 when this voltage is within 4V of the half-supply value of the circuit. In split-supply circuits, of course, the 'half-supply' value equals zero volts. When pin 1 is at the half-supply value, the output amplitude of the pin 2 signal is approximately zero. When the pin 1 voltage is increased above the half-supply value the pin 2 signal rises in direct proportion. When the pin 1 voltage is reduced below the half-supply value the pin 2 signal again rises in direct proportion, but the phase of the output signal is reversed. This last-mentioned phenomenon can be utilised for phase-shift keyed (PSK) and suppressed carrier a.m. generation.

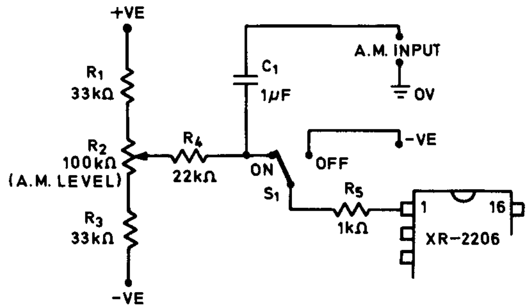


Fig. 3.17. Add-on a.m. facility

The pin 1 terminal of the i.c. can also be used to facilitate gate-keying or pulsing of the pin 2 output signal. This can be achieved by biasing pin 1 to approximately half-supply volts to give zero output at pin 2, and then imposing the gate or pulse signal on pin 1 to raise the pin 2 signal to the required turn-on amplitude. The total dynamic range of amplitude modulation is approximately 55dB.

Frequency sweep and f.m. generation

The frequency of oscillation of the XR-2206 is proportional to the total timing current (I_T) drawn from pin 7 or 8, and is given by

$$f = \frac{320 \times I_T \text{ (mA)}}{C(\mu F)} \text{ Hz.}$$

The timing terminals (pins 7 or 8) are low-impedance points, and are internally biased at +3V with respect to pin 12. The frequency varies linearly with I_T over the approximate current range 1μA to 3mA. Consequently, the frequency can be varied by (a) wiring a variable current-determining resistor between pin 12 and the timing terminal, or (b) by applying a variable voltage in the range 0 to +3V between pin 12 and the timing terminal via a current-limiting resistor, or (c) by a combination of these two techniques. The technique outlined in (b) can be used to frequency-sweep the output signals of the XR-2206, and the technique mentioned in (c) can be used to frequency-modulate the output signals.

Fig. 3.18 shows the basic connections of a simple frequency-sweep circuit with a 6:1 range of frequency coverage. Here, a sawtooth frequency-sweep signal, which has a peak amplitude of 2.5V, is applied between pin 12 and the pin 7 timing terminal via limiting resistor R_1.

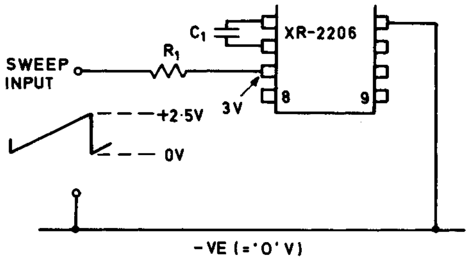


Fig. 3.18. Frequency-sweep circuit gives 6:1 frequency range

Consequently, when the instantaneous peak value of the sawtooth voltage is zero, 3V is developed across R_1, and the frequency is $1/RC$ Hz, as in the case of a simple resistance-controlled XR-2206 oscillator. When, on the other hand, the instantaneous peak value of the sawtooth voltage is 2.5V, only 0.5V is developed across R_1, and the R_1 current is only 1/6th of that of the case we have just looked at, so the frequency falls to $1/6RC$ Hz. The frequency of oscillation is thus determined by the instantaneous value of the sweep voltage and by the R_1 and C_1 values. The frequency can, in theory, be varied over a range of at least 1000:1 by using this simple frequency-sweep technique.

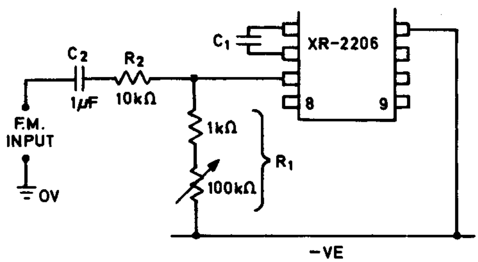


Fig. 3.19. Simple f.m. facility for the XR-2206

Fig. 3.19 shows the essential connections of a simple f.m. facility for the XR-2206 i.c. Here, in the absence of an f.m. input signal, the operating frequency is determined by R_1 and C_1, as in the case of a conventional XR-2206 oscillator. When the f.m. input signal is connected to the circuit, the f.m. input currents are added to those of R_1

via R_2, so the timing currents and thus the operating frequency of the circuit are effectively modulated by this input signal. C_2 is used to block any d.c. components between the input signal and the main timing resistor.

A weakness of the simple Fig. 3.19 circuit is that, for a given amplitude of input signal, the percentage deviation or sensitivity of the f.m. facility varies with the setting of the R_1 frequency control. This snag

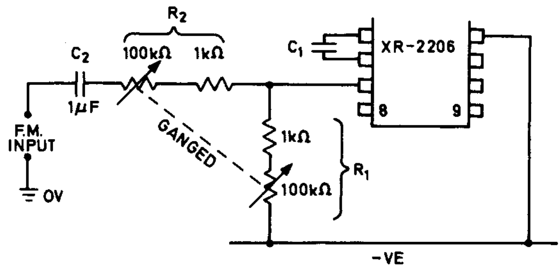


Fig. 3.20. Constant-sensitivity f.m. facility

is overcome in the constant-sensitivity circuit of Fig. 3.20. Here, a dual ganged pot is used with one arm connected to the R_1 frequency-determining network and the other arm connected to the R_2 f.m.-sensitivity network, so that the sensitivity is automatically adjusted to track with the frequency setting.

It should be noted that the frequency-sweep and f.m. generation facilities control both the pin 2 and pin 11 output terminals of the XR-2206 i.c.

Power supplies for the XR-2206 i.c.

Finally, to conclude this chapter, a few notes on choosing power supplies for the XR-2206 i.c. It should be noted that the XR-2206 i.c. has built-in voltage regulation circuitry controlling its v.c.o. section, and that the amplitude of the pin 2 output signal is not significantly influenced by variations in circuit supply voltage. Consequently, there is little advantage in using higher supply voltages than are absolutely necessary for correct operation of the XR-2206. In most cases, single-ended or split 12V supplies are adequate.

Fig. 3.21 shows the practical circuit of an a.c. powered zener-regulated split-supply 12V (6V–0–6V) circuit suitable for powering the i.c., and Fig. 3.22 shows the alternative connections for using a battery power supply. The XR-2206 draws a typical current of about 15mA when used with these supplies.

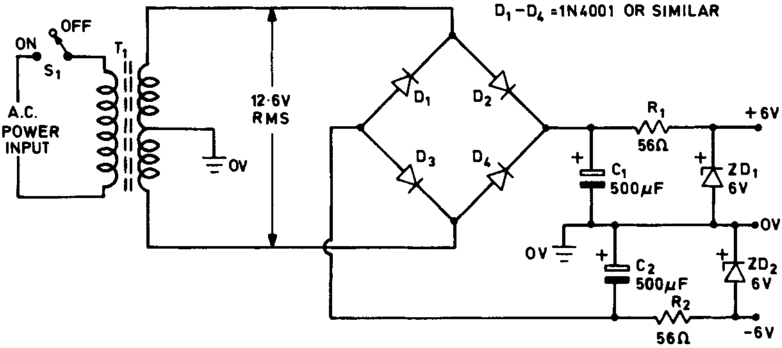


Fig. 3.21. Zener-regulated 6-0-6V supply for the XR-2206

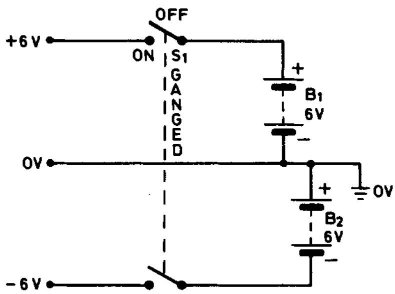


Fig. 3.22. 6-0-6V battery supply for the XR-2206

CHAPTER 4

TYPE LM380 2-WATT AUDIO AMPLIFIER CIRCUITS

The LM380 integrated circuit from National Semiconductors is an easy-to-use general-purpose power audio amplifier. It has an internally fixed voltage gain of 50 (34dB), a nominal output power capability of 2W r.m.s., and an output that is well protected with short circuit limiting and thermal shutdown circuitry. The device has a typical power bandwidth of 100kHz at 2W output into an 8Ω load, typically produces distortion of only 0.2% at 2W into an 8Ω load, and can be used with any single-ended supply voltages in the range 8V to 22V.

This chapter briefly describes the operating characteristics of the LM380 i.c., and shows ten simple but practical applications of the device.

How it works

Fig. 4.1a shows the basic internal circuit of the LM380 i.c. Here, Q_1 and Q_2 are wired as pnp emitter followers that drive the Q_3–Q_4 differential amplifier pair of pnp transistors: pnp transistors are used in these stages so that input signals can be d.c. referenced to the ground line, thus enabling input transducers to be directly connected between the ground and input lines. The output of the differential amplifier stage is direct coupled into the base of Q_{12}, which is wired as a simple common emitter amplifier with Q_{11} acting as its high impedance (constant-current) collector load, and the collector signal of Q_{12} is fed to the output terminal of the i.c. via the Q_7–Q_8–Q_9 quasi-complementary emitter follower set of output transistors. The output currents of Q_7 and Q_9 are rated at 1.3A peak.

Bias-determining and gain controlling resistor networks are built into the LM380 i.c. Feedback resistor R_2 is wired between the output terminal of the i.c. and one emitter side of the differential amplifier

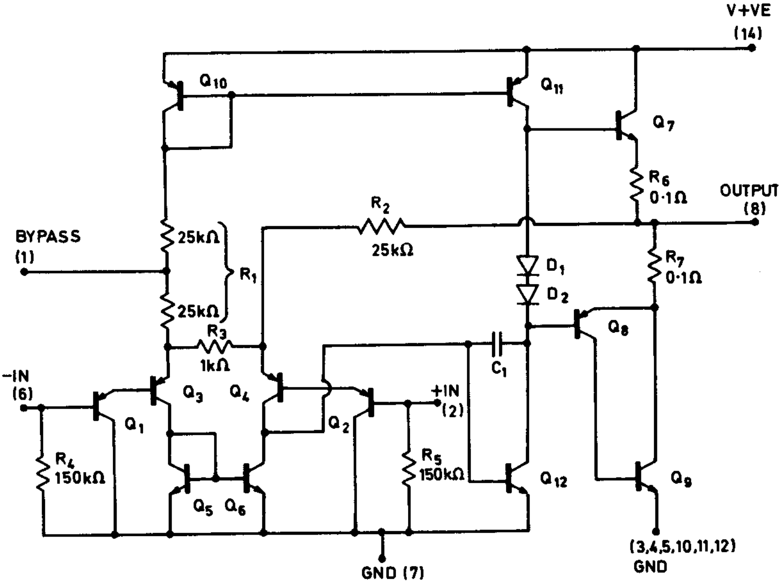


Fig. 4.1a. Basic internal circuit of the LM380 i.c.

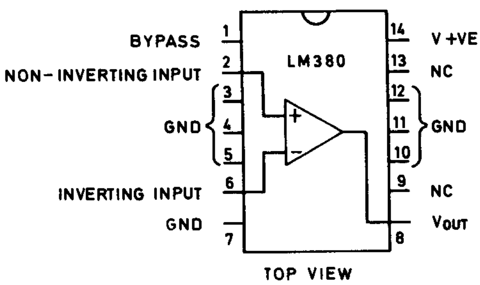


Fig. 4.1b. Outline and pin connections of the standard 14-pin d.i.l. (d.i.p.) version of the LM380 i.c.

stage, and has half the value of R_1; the action of these two resistors is such that the amplifier output automatically balances at a quiescent potential of approximately half supply-line voltage. The voltage gain of the i.c. is internally fixed at x50 or 34dB by the ratios of R_2 and R_3.

Fig. 4.1b shows the outline and pin connections of the LM380 i.c., which is housed in a standard 14-pin d.i.l. package. The package contains a copper lead frame that acts as a heat sink and is internally connected to the three centre pins on either side of the i.c. (pins, 3, 4, 5, 10, 11, and 12). This frame enables the device to support 1.1W dissipation at 50°C ambient, or 1.5W at 25°C ambient. Device dissipation can be increased to 3W at 50°C ambient or 3.7W at 25°C ambient by soldering the six heat sink pins into a printed circuit board with 6 square inches (37.5 square cm) of standard copper foil. With an infinite heat sink the device rating rises to 8W at 50°C or 10.5W at 25°C.

The LM380 i.c. is a very easy device to use. Input signals can be direct coupled to either the inverting (pin 6) or non-inverting (pin 2) input terminals, which have input impedances of about 150kΩ. An unused input terminal can either be left floating, or can be shorted directly to ground, or can be tied to ground via a resistance. The output speaker load is connected between the pin 8 output terminal and ground (pin 7) via a large-value electrolytic blocking capacitor in most applications.

The LM380 i.c. can be used with any single-ended supply voltages in the range 8V to 22V. Normal power supply decoupling precautions must be taken when using the i.c. Specifically, the supply must be decoupled via a 47μF or larger electrolytic capacitor located close to the i.c., and a 0.1μF ceramic capacitor must be wired as near as possible between the pin 14 and ground terminals of the i.c., as a precaution against parasitic oscillation. If the i.c. is to be used to drive low-impedance inductive loads, such as loud speakers, a Zobel network (comprising a 2.7Ω resistor and a 0.1μF capacitor in series) must be wired between the output (pin 8) and ground (pin 7 or 3) terminals of the i.c., to protect the circuit against high frequency oscillations. If the power supply lines to the i.c. have high ripple, a 10μF or greater electrolytic should be wired between the *bypass* pin 1 and ground terminals of the i.c. to prevent this ripple reaching the speaker. The i.c. gives greater than 37dB ripple rejection at 50Hz and above when fitted with this 10μF capacitor.

Practical applications

Fig. 4.2 shows how the LM380 can be used as a simple non-inverting 2W amplifier. Here, the input signal is direct coupled between ground and the pin 2 non-inverting input terminal of the i.c., which has an input impedance of about 150kΩ. The 8Ω speaker is connected between ground and the pin 8 output terminal of the i.c. via 470μF blocking capacitor C_3. Zobel network R_1–C_4 is connected across the output of the i.c. to inhibit oscillations that might otherwise occur when using low-impedance inductive speaker loads. Supply decoupling capacitors C_1 and C_2 protect the i.c. against parasitic oscillation.

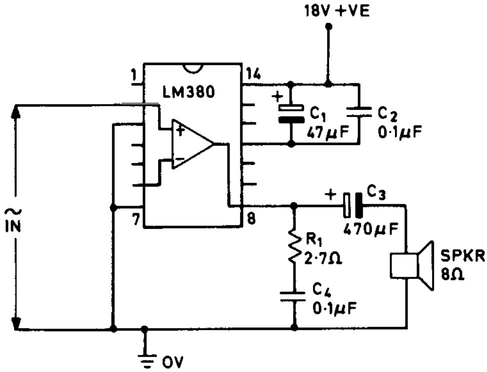


Fig. 4.2. Simple non-inverting 2W amplifier

The LM380 can be used in the inverting amplifier mode by simply connecting the input signal directly to the pin 6 inverting terminal of the i.c. and leaving the unused pin 2 terminal floating, as shown in

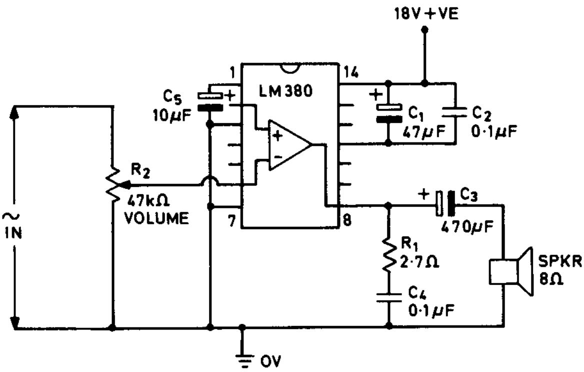


Fig. 4.3. Inverting 2W amplifier, with volume control and ripple rejection

Fig. 4.3. This diagram also shows how a potential divider volume control facility (R_2) can be added to the circuit, and how improved supply-line ripple rejection can be obtained by wiring C_5 between bypass pin 1 and ground.

Fig. 4.4 shows the circuit of a simple 2W phono amplifier, which can be used with a ceramic or crystal pick-up cartridge. Here, R_2 and R_3 act as a simple volume control, and C_5 and R_4 act as a simple tone control that gives variable high-frequency roll-off. R_3 is wired in series

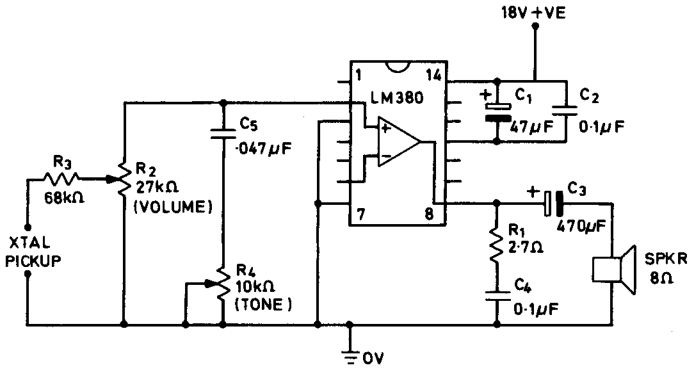


Fig. 4.4. Simple 2W phono amplifier

between the pick-up and the R_2 slider, to give the circuit a reasonably high input impedance. A minor weakness of this simple circuit is that, because of the presence of R_3, signal attenuation occurs between the pick-up and the input pin of the LM380 at all settings of R_2.

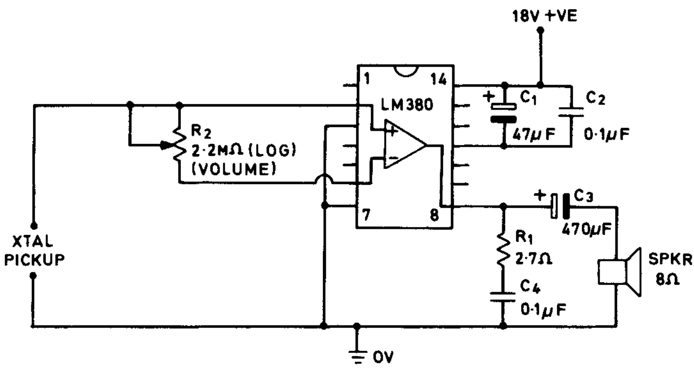


Fig. 4.5. 2W amplifier with 'common mode' volume control

An alternative, and possibly better, system of volume control is shown in the circuit of Fig. 4.5. Here, the pick-up signal is fed directly to the non-inverting (pin 2) terminal of the LM380, and a high-value variable resistance is wired between the inverting (pin 6) and non-inverting terminals of the i.c. Remembering that each of these terminals has an input impedance of 150kΩ, and that the i.c. acts as a differential amplifier when both terminals are in use, it can be seen that if R_2 has a value of zero ohms identical input signals are applied to each terminal, thus producing zero output voltage, and that under this condition the input impedance of the circuit is equal to 150kΩ/2.

When, on the other hand, R_2 is set at maximum value, all of the input signal is applied to pin 2 of the i.c. and only a negligible part of the signal reaches pin 6, so the voltage gain of the circuit is close to its full single-ended value and the i.c. produces a high output voltage, and the input impedance of the circuit is close to the 150kΩ value of pin 2. R_2 thus acts as an effective 'common mode' volume control, and does not appreciably detract from the high input impedance characteristics of the LM380 i.c.

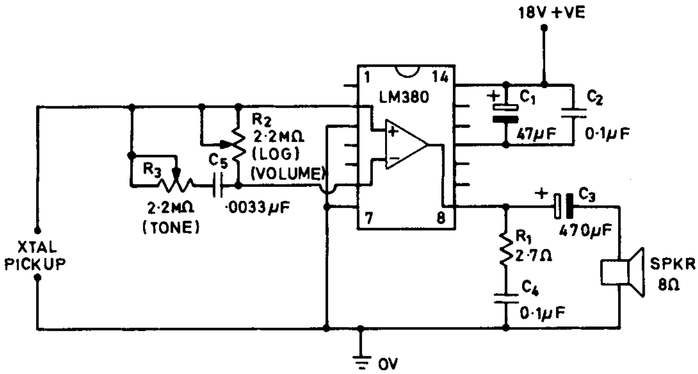


Fig. 4.6. Practical 2W phono amplifier with 'common mode' volume and tone controls

A 'common mode' tone control can be combined with the 'common mode' volume control of Fig. 4.5 by using the connections shown in Fig. 4.6. Here, tone control C_5–R_3 acts as a simple top-cut variable filter that has a turn-over frequency of about 100Hz. This circuit acts as an excellent general-purpose low-cost phono amplifier.

Fig. 4.7 shows how the above circuit can be adapted for use as a stereo amplifier. Points to note here are that ganged volume (R_2) and tone (R_3) controls are used, and that only a single high-value supply decoupling capacitor (C_1) is used. If this amplifier suffers from 'hum' when used in mains (line) powered applications, each i.c. should be provided with ripple rejection by wiring a 10µF electrolytic capacitor between pin 1 of the i.c. and ground.

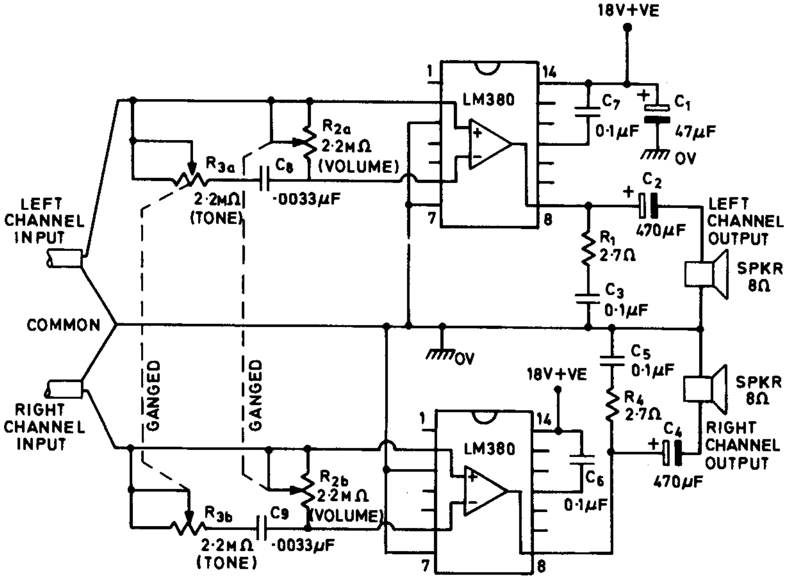


Fig. 4.7. Simple stereo phono-amplifier gives 2W per channel

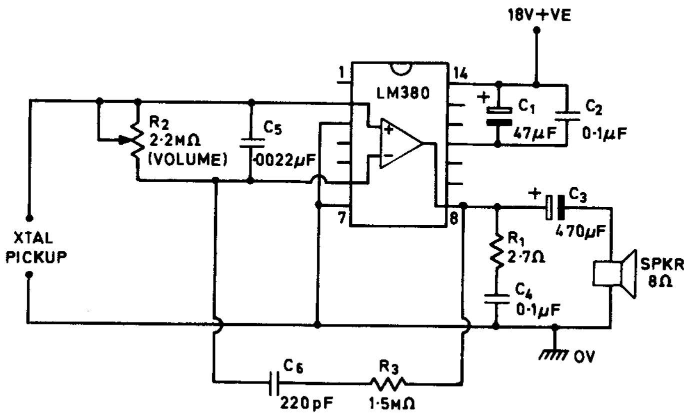


Fig. 4.8. RIAA phono amplifier

In some phono amplifier applications the user may prefer to provide the amplifier with fixed RIAA equalisation characteristics, rather than with a variable tone control. In such cases, the circuit of Fig. 4.8 can be used.

The LM380 i.c. has plenty of uses other than in phono applications. Fig. 4.9 shows how the i.c. can be used in conjunction with a field-effect transistor (f.e.t.) to make a high-input-impedance (10MΩ) general-purpose 2W amplifier. Here, the f.e.t. (Q_1) is wired as a simple source

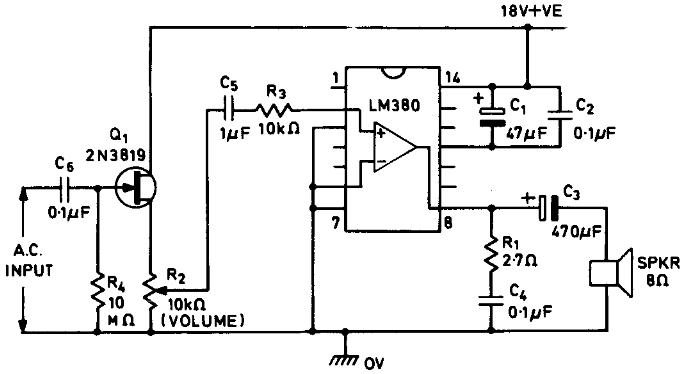


Fig. 4.9. High imput impedance (10MΩ) general purpose 2W amplifier

follower, with its input impedance determined by R_4, and the output of the f.e.t. is fed to the input of the i.c. via volume control R_2. Resistor R_3 is wired in series with the input of the i.c. to prevent switch-on transients from damaging the device.

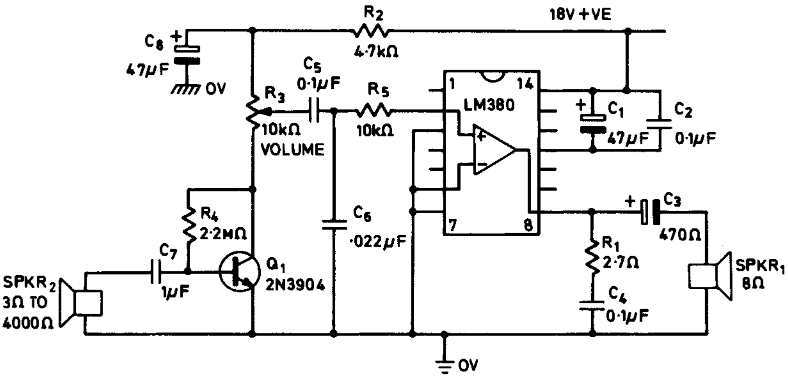


Fig. 4.10. Baby alarm

Fig. 4.10 shows how the LM380 can be used in conjunction with a single bipolar transistor to make a useful and inexpensive baby alarm. Here, the sound of the baby is picked up by *SPKR*$_2$, which can be any speaker or microphone with an impedance in the range 3Ω to 4000Ω, and this signal is passed to the base of transistor Q_1. Q_1 is wired as a simple common emitter amplifier, with its collector load R_3 also acting as a volume control, and the output of R_3 is passed on to the input of the i.c. via C_5 and R_5. C_6 is used as a basic filter that attenuates high frequency signals, and R_2 and C_8 provide a decoupled supply to the Q_1 amplifier stage.

Finally, to complete this chapter, Fig. 4.11 shows how a pair of LM380 i.c.s can be interconnected in the bridge configuration to give

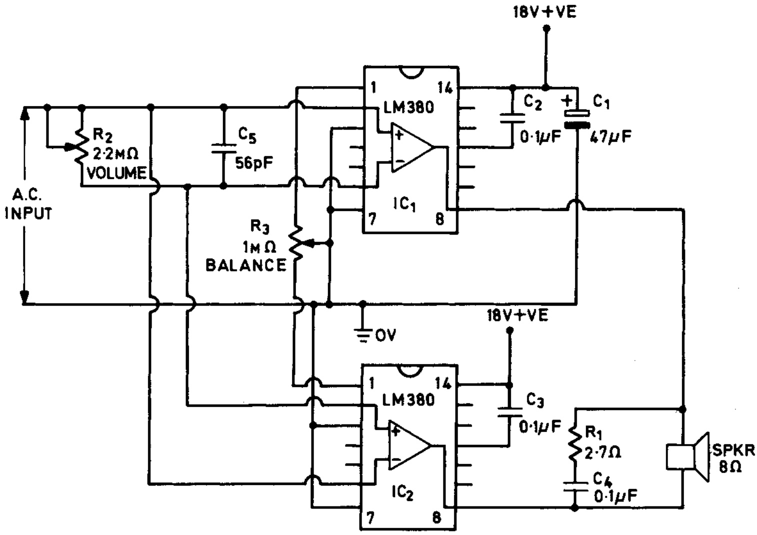


Fig. 4.11. 4W bridge-configuration amplifier using common-mode volume control

an output power roughly double that of a single i.c. Here, the input signal is applied to the non-inverting input terminal of one amplifier and to the inverting terminal of the other, thus producing antiphase output signals from the two i.c.s. and hence dividing the output power dissipation equally between the two i.c.s. The Fig. 4.11 circuit is shown with a 'common mode' volume control, but this bridge configuration can be used equally well with potential divider type volume controls.

Note in the Fig. 4.11 circuit that the speaker is direct coupled between the two output pins (pins 8) of the i.c.s. The basic quiescent output voltage of the LM380 is specified as between 8V and 10V when used with an 18V supply, so a quiescent differential of 2V d.c. can exist across the speaker load of the Fig. 4.11 circuit, causing a quiescent current up to 250mA to flow in the speaker. Balance control R_3 is included in the circuit to enable these differential voltages and currents to be reduced to zero. To set up the circuit, simply wire a current monitor in series with the speaker and, with zero input signal applied, adjust R_3 to give zero current reading in the monitor. All adjustments are then complete.

CHAPTER 5

TYPE 723 VOLTAGE REGULATOR CIRCUITS

The type 723 voltage regulator i.c. is a versatile and readily available device that can be used in a variety of fixed or variable voltage power supply applications. It can be operated with any supply voltage in the range 9.5V to 40V, and can provide well regulated outputs in the range 2V to 37V. The device incorporates current-limiting circuity which can be programmed via a single external resistor, and the i.c. can supply currents up to 150mA directly or up to tens of amps when used in conjunction with external transistors.

The type 723 i.c: How it works

The type 723 voltage regular i.c. is produced by several different manufacturers and is available with a variety of i.c. codings, but is generally known simply as a '723 voltage regulator'. Fig. 5.1a shows the equivalent circuit of the i.c., and Figs. 5.1b and 5.1c show the outlines and pin notations of the two most popular forms of packaging of the device. Throughout this chapter, all practical regulator circuits are based on the 14-pin d.i.l. version of the i.c.

The 723 i.c. contains fifteen bipolar transistors, one f.e.t., three zener diodes, twelve resistors, and one capacitor. As shown in the equivalent circuit of Fig. 5.1a, these components are arranged in the form of a built-in voltage reference, an error amplifier, a series-pass output transistor, and a current-limiting transistor. The voltage reference section is temperature compensated, produces a nominal output of 7.2V, and can provide up to 15mA of output current.

The built-in error amplifier of the 723 i.c. acts like an operational amplifier. It has an inverting and a non-inverting input terminal, and its output is used to directly drive the base of the series pass output

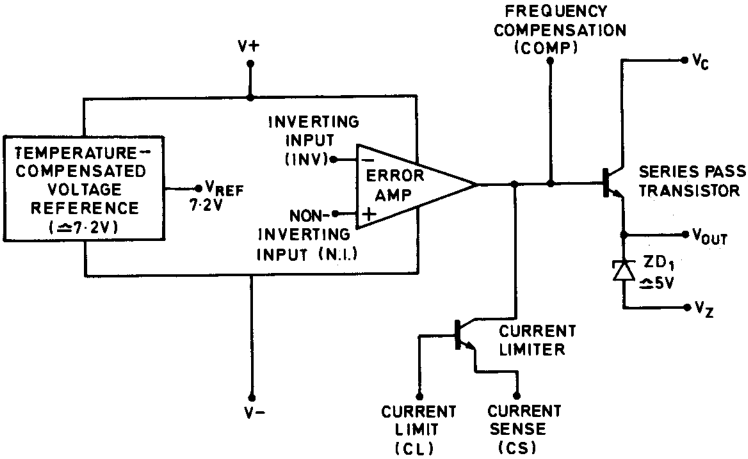


Fig. 5.1a. Equivalent circuit of the type 723 voltage regulator i.c.

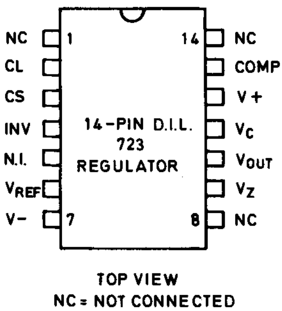


Fig. 5.1b. Outline and pin notations of the 14-pin d.i.l. (d.i.p.) version of the 723 i.c.

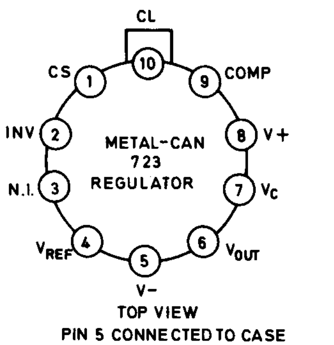


Fig. 5.1c. Outline and pin notations of the metal can version of the 723 i.c.

transistor. Fig. 5.2a shows how the basic elements of the i.c. can be interconnected to form a simple 2V to 7.2V voltage regulator. Here, the output of the series pass transistor is connected directly to the inverting input of the error amplifier, so that the two elements act as a precision voltage follower that gives unity gain between the non-inverting input terminal and the inverting feedback loop. The input to the non-inverting terminal is derived from the built-in voltage reference via the R_1–R_2

potential divider, and can be varied between 2V and 7.2V by choice of R_1 and R_2 values. Thus, the output voltage of the circuit is the same as the voltage on the non-inverting input terminal of the error amplifier, and approximates

$$7.2\text{V} \times R_2/(R_1 + R_2).$$

The elements of the 723 i.c. can be wired to give regulated outputs in the range 7.2V to 37V by using the basic connections shown in Fig. 5.2b. Here, the 7.2V reference is fed directly to the non-inverting

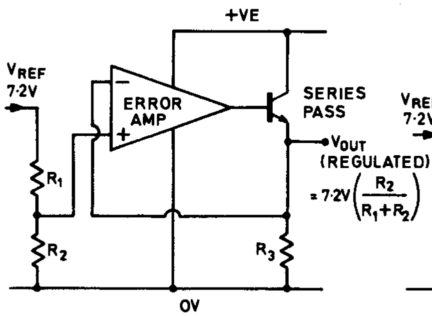


Fig. 5.2a. Basic 2–7.2V regulator

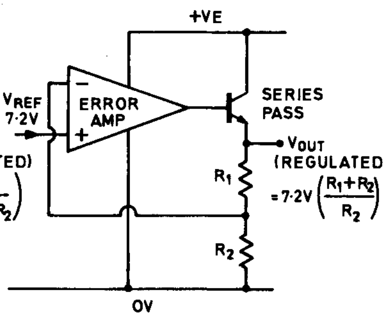


Fig. 5.2b. Basic 7.2–37V regulator

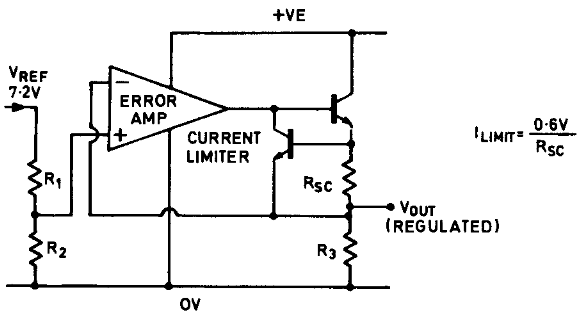


Fig. 5.2c. Current limiting facility applied to the basic regulator circuit

terminal of the error amplifier, and the error amplifier and series pass transistor are wired as a composite non-inverting amplifier with gain controlled by the ratios of R_1 and R_2. Thus, the regulator output voltage is a precise multiple of the input reference voltage, and approximates

$$7.2\text{V} \times (R_1 + R_2/R_2)$$

Finally, Fig. 5.2c shows the basic method of using the current limiting facility of the i.c. Here, current-sensing resistor R_{sc} is wired in series with the output terminal of the i.c. so that the voltage across R_{sc} is directly proportional to the output current, and the base-emitter junction of the current limiter transistor is wired across this sensing resistor. Normally, the current limiting transistor is cut off and has no effect on the circuit, but when a short is placed across the output of the circuit the output current rises to such a level that the current limiter is driven close to saturation by the R_{sc} voltage, and thus reduces the emitter-base drive voltage of the series output transistor and hence limits the output current. Heavy negative feedback is involved in this action, and the maximum output current is automatically limited to a value of V_{eb}/R_{sc}, or approximately $0.6V/R_{sc}$. Thus, an R_{sc} value of 100Ω gives a limit current of about 6mA, and an R_{sc} value of 10Ω gives a limit current of 60mA. This current limiting facility can be used in conjunction with sensing resistors driven by external high-current output transistors, and can thus provide current limiting at values up to tens of amps.

Note that, since R_{sc} is wired into the negative feedback loop of the circuit, it has negligible effect on the output impedance of the regulatoı.

Using the 723 voltage regulator i.c.

Basic points to note when using the 723 voltage regulator i.c. are as follows.

(1) The i.c. can be powered from any supply voltage in the range 9.5V to 40V. The supply voltage must, however, always be at least 3V greater than the desired regulator output voltage.

(2) The regulated output voltage can be set at any value in the range 2V to 37V.

(3) The maximum current that can safely be taken directly from the output of the i.c. is 150mA, but in most cases the 'safe' current limit is determined by the power rating limitations of the i.c., and equals P_{max}/V_{supply} under shorted-output conditions.

(4) The internal power dissipation limit of the plastic 14-pin d.i.l. version of the 723 i.c. is 660mW at an ambient temperature of 25°C, and is likely to be the major factor limiting the maximum safe output current that can be taken from the device. As an example, the maximum

safe current that can be taken from a regulator circuit powered from a 15 volt supply is 660mW/15V = 44mA. The output capability can easily be boosted by using the regulator in conjunction with an external power transistor stage.

(5) The standby current drawn by the i.c. is typically 1.3mA when powered from a 30V supply, and is not greatly influenced by variations in supply voltage.

(6) The maximum voltage that can be safely applied to either input terminal of the error amplifier of the i.c. is 7.5V.

(7) A zener diode is wired between the V_{out} and V_z terminals of the i.c. This diode gives a zener voltage of about 6V, and can provide maximum currents up to 25mA.

(8) In all 723 regulator circuits an external parasitic-stopping capacitor must be wired to the *comp* terminal of the i.c. The value and connections of this capacitor are determined by the circuit configuration that is used.

(9) The built-in voltage reference of the i.c. is temperature compensated. It has typical and worse-case temperature coefficients of 0.003%/°C and 0.015%/°C respectively.

(10) The current limiting facility of the i.c. is temperature sensitive, and has a first-order coefficient of about –0.3%/°C. Thus, if the current is set to limit at 100mA at 25°C, it will be found that limiting occurs at about 92.5mA at 50°C and at about 107.5mA at 0°C.

(11) The 723 regulator i.c. has a tendency to become unstable when feeding reactive (rather than purely resistive) loads. To enhance circuit stability and minimise the regulator output impedance a large smoothing capacitor should be wired across the protected output of the 723 regulator circuit. To protect the i.c. against damage from stored energy at the moment of power switch-off, a safety diode should be connected between the positive supply line and the protected output of the regulator circuit so that the diode is normally reverse biased but conducts under the above condition.

Practical 723 voltage regulator circuits

In all practical circuits shown throughout the remainder of this chapter the designs are presented in both (a) conventional and (b) constructional form. The 'conventional' presentation aids the understanding of circuit operation. The 'constructional' presentation is a practical wiring aid, and shows the circuit redrawn around the standard 14-pin d.i.l. outline of the i.c.

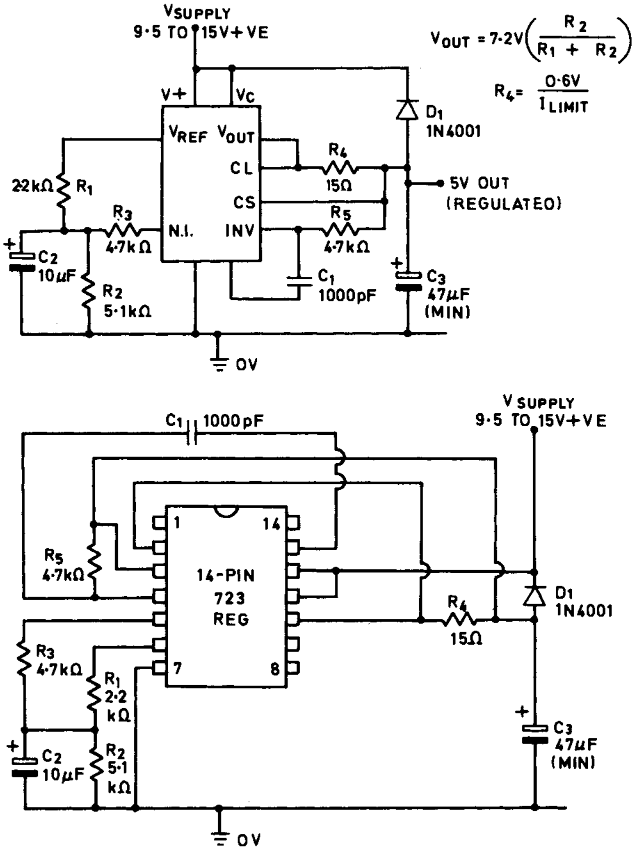


Fig. 5.3. Practical low-voltage (2–7.2V) regulator gives protected 5V output at 40 mA maximum

Fig. 5.3 shows the circuit and connections of a practical low-voltage (2V to 7.2V) regulator with preset current limiting. With the component values shown the circuit is designed to produce an output of 5V at limit currents up to 40mA.

Here, the built-in reference voltage is fed to the non-inverting terminal of the error amplifier via the R_1–R_2 potential divider (which produces a stable output of 5V) and via thermal compensation resistor R_3, and the error amplifier and built-in series-pass transistor are wired as a unity-gain voltage follower network, which thus produces an output of 5V. Current limiting is provided via R_4, which is included in the negative feedback stage of the voltage follower network.

The output voltage of the above circuit is controlled by R_1 and R_2, and approximates 7.2V x $R_2/(R_1 + R_2)$. The voltage can be varied over the range 2V to 7.2V by selection of resistor values. Resistor R_5 provides impedance isolation between the *INV* and *CS* terminals of the i.c. and enhances circuit stability, and R_3 gives thermal compensation to the *NI* terminal of the regulator. C_3 is wired across the output of the regulator, and serves the dual function of enhancing circuit stability and giving a low high-frequency output impedance to the regulator circuit. Component D_1 acts as a safety diode and protects the i.c. against damage from stored energy at the moment of power switch-off. Current limiting is controlled by R_4, which has a value determined on the basis of $0.6V/I_{limit}$. Thus, for a 40mA limit current R_4 requires a value of 15Ω.

The Fig. 5.3 circuit can be usefully modified in a number of way. It can be made to produce variable output voltages by using a variable potential divider in place of R_1 and R_2, and its current drive capability can be increased by taking the output via an external power transistor. Both of these modifications are shown in Fig. 5.4.

Here, the reference voltage (and thus the output voltage) are made variable over the approximate range 2V to 7.2V via the R_1 and R_2 variable potential divider, and the current drive capability of the regulator is boosted via series-pass transistor Q_1. This transistor, which must be mounted on a heat sink, increases the output current drive capability of the circuit by a factor equal to the transistor h_{fe} ($h_{fe\ min}$ = 20 for the 2N3055 transistor). In the diagram, current-sensing resistor R_4 is given a value of 1Ω, thus giving a short-circuit current value of about 600mA. Thus, under worst-case (15V supply voltage and a Q_1 h_{fe} of 20) short-circuit conditions the power dissipation of the 723 regulator i.c. is restricted to a safe value of about 400mW.

Fig. 5.5 shows how the current-drive capabilities of the above circuit can be further boosted by replacing the external series-pass transistor with a Darlington or super-alpha connected pair of transistors. Both transistors must be mounted on heat sinks. In this circuit the current drive capability of the regulator is increased by a factor equal to the h_{fe} product of the two transistors, i.e. if both transistors have h_{fe} values of 20, the drive capability is increased by a factor of 400, thus enabling output currents up to 12A to be taken while the regulator i.c. is providing only 30mA and dissipating worst-case power of less than 400mW. In the actual circuit of Fig. 5.5 the short-circuit output current is restricted to about 2A via 0.3Ω current-sensing resistor R_4, but in practice the current capability can be increased to a maximum of about 12A by simply reducing the R_4 value.

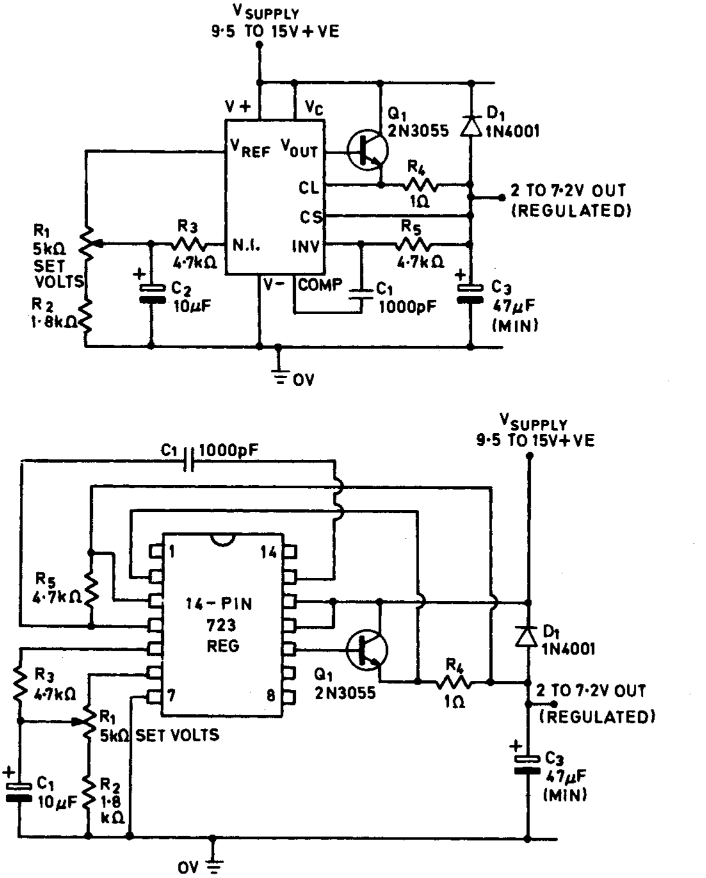


Fig. 5.4. Variable low-voltage (2–7.2V) regulator gives protected 600mA output

Note in the Fig. 5.5 circuit that the feedback lead to the *INV* terminal of the i.c. is taken from the positive output terminal of the design, rather than directly from the *CS* terminal, so that the output lead is included in the negative feedback loop of the circuit and its effective impedance is consequently reduced to a negligible value. Connecting leads have typical resistances of about 20mΩ per metre, thus giving volt drops of about 20mV/metre/A, and a substantial difference can thus exist between the output and the load voltages of a system if the lead is not remotely included in the feedback network of the regulator.

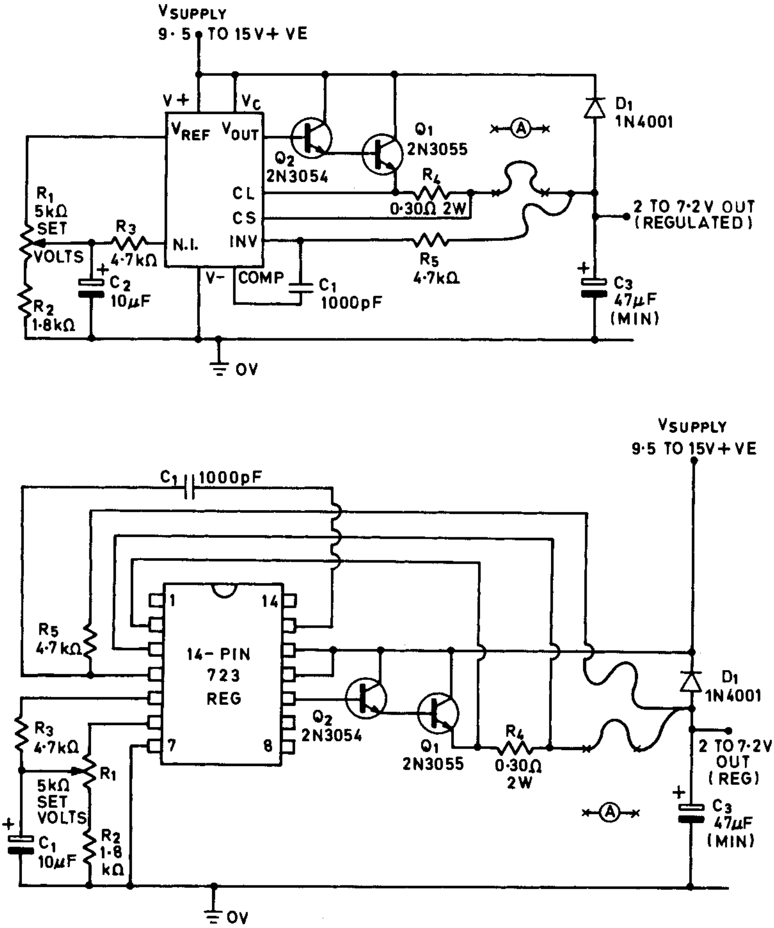


Fig. 5.5. Variable low-voltage (2–7.2V) regulator gives protected 2A output with remote feedback facility. Circuit can be modified to provide output currents up to 12A

Also note that a current monitor meter can be wired into the output of the circuit at point 'x-x' without significantly increasing the output impedance of the circuit, since this meter is also included in the negative feedback loop of the design.

Fig. 5.6 shows how the 723 regulator i.c. can be used to generate voltages in the range 7V to 37V. Here, the direct output of the built-in 7.2V voltage reference is fed to one side of the unit's voltage comparator, and the comparator and the integrated series-pass transistor are wired as

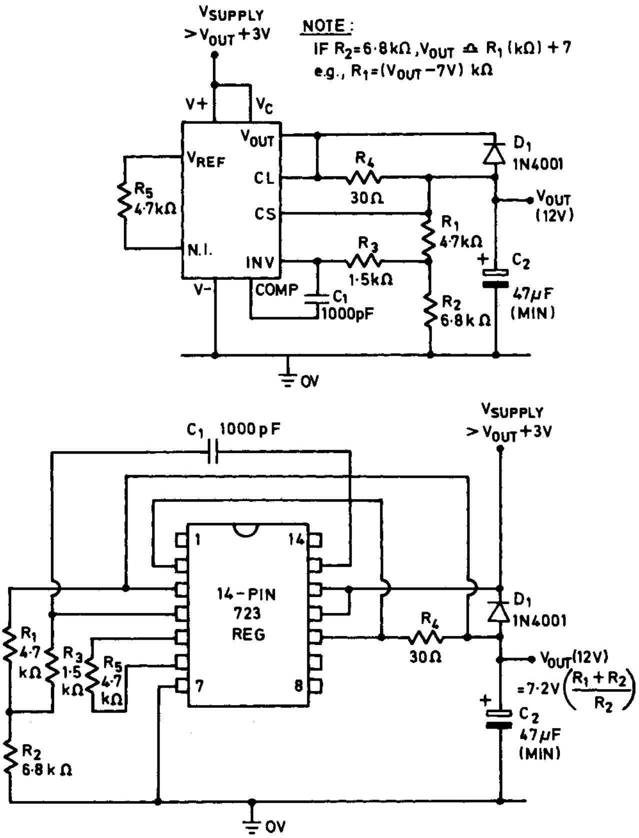


Fig. 5.6. Simple high-voltage (7–37V) regulator or precision variable zener reference: with the component values shown, the circuit gives an output of 12V and has a limit (short circuit) current of 20mA

a composite non-inverting d.c. amplifier with its gain controlled by the ratios of R_1 and R_2. Thus, the output of the circuit equals 7.2V x (R_1 + R_2/R_2). Note that if R_2 is given a value of 6.8kΩ the output voltage then approximates R_1 + 7 when the R_1 value is given in kΩ, e.g., R_1 (in kΩ) approximately equals V_{out} – 7. Thus, for an output of 12V R_1 needs a value of 5kΩ, and for an output of 15V R_1 needs a value of 8kΩ. With the component values shown, the Fig. 5.6 circuit gives an output of about 12V.

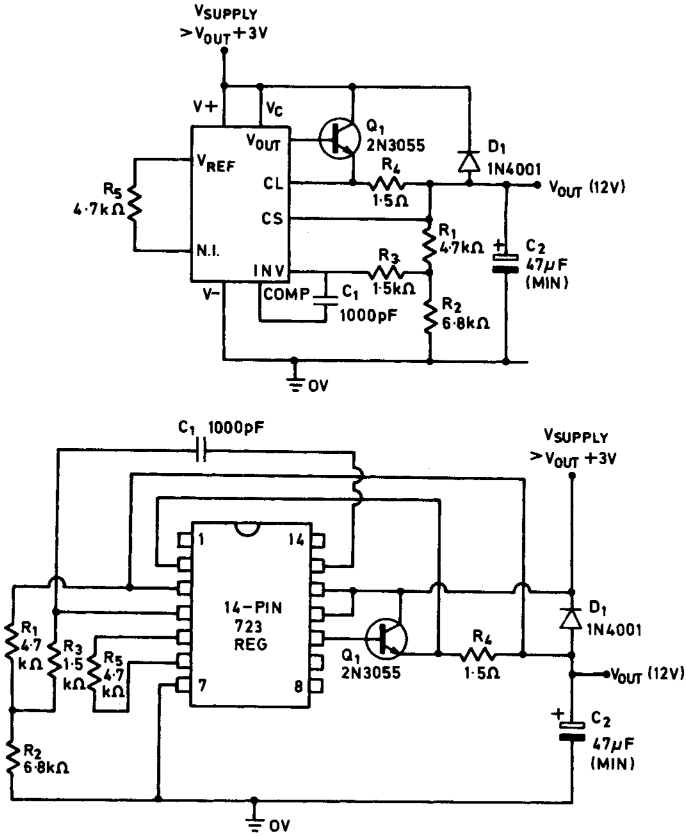


Fig. 5.7. Boosted-output high-voltage (7–37V) regulator gives 12V output at 400mA

The supply voltage used with the Fig. 5.6 circuit must be at least 3V greater than the required output voltage of the regulator. The maximum power dissipation of the 723 i.c. is limited to 660mW at an ambient temperature of 25°C, and the safe (short-circuit) output current of the i.c. is thus determined on the basis of 660mW/V_{supply}. Thus, the 'safe' output is restricted to about 22mA with a 30V supply, or to about 44mA with a 15V supply. With the component values shows, the short-circuit output current of the Fig. 5.6 circuit is restricted to 20mA by R_4, and the circuit thus acts as only a low current voltage regulator. The circuit can thus be used as a precision variable zener reference, since it has a very low temperature coefficient and can be set to give any reference voltage in the range 7V to 37V.

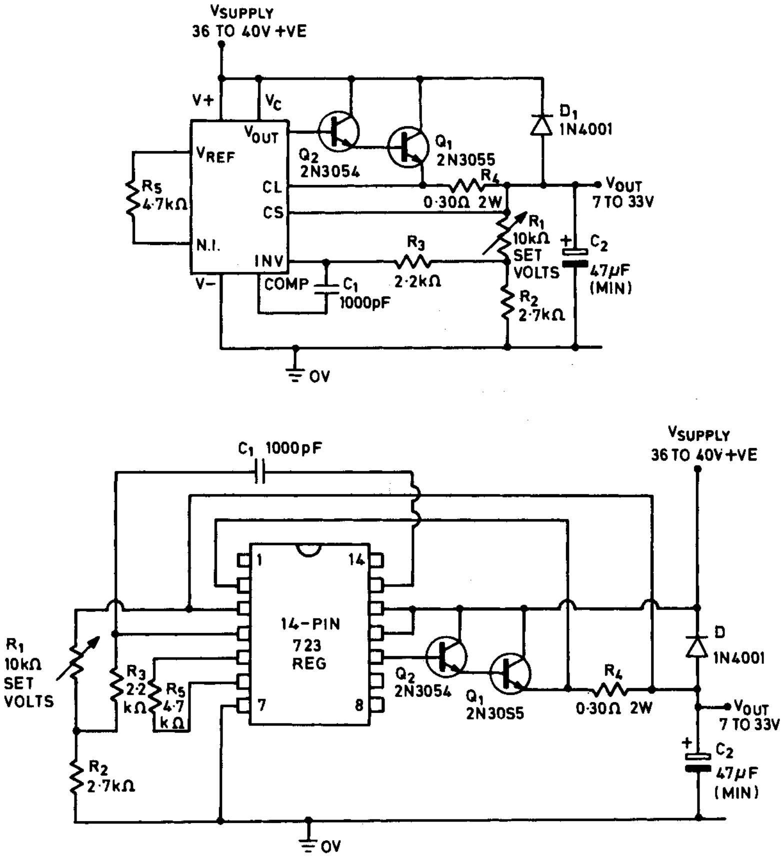


Fig. 5.8. Variable output (7–33V) regulator gives 2A limit current

The output current capability of the basic Fig. 5.6 circuit can be boosted by connecting an external booster transistor in series with the output of the regulator, as shown in Fig. 5.7. Here, a 2N3055 series pass booster transistor is used, and boosts the available output current by a factor of 20, to 400mA. Current-sensing resistor R_4 is reduced in value to 1.5Ω, to ensure a short-circuit limit current of 400mA.

The output current capability of the above circuit can be further boosted by using a Darlington or super-alpha connected pair of transistors in place of a single series-pass external transistor, as shown in Fig. 5.8. Here, the short-circuit current is set at 2A via R_4. This circuit also shows how the output voltage of the regulator can be made fully variable from 7V to 33V by using a 10kΩ variable resistor in the R_1 position.

The operating theory of the variable-voltage facility of the Fig. 5.8 circuit is quite simple. The output voltage of the regulator circuit equals

$$V_{\text{ref}}\left\{\frac{R_1 + R_2}{R_2}\right\}$$

V_{ref} is the built-in reference voltage (about 7V) of the i.c., and $(R_1 + R_2/R_2)$ represents the voltage gain of the amplifier section of the circuit. Thus, when R_1 is set to zero the voltage gain of the circuit equals unity and the output voltage equals 7V, but when R_1 is set at its maximum value the voltage gain equals 4.7 and the output voltage equals 33V. Note that the minimum output voltage of the circuit is determined by the reference voltage fed to the *NI* terminal of the i.c., and that the maximum voltage is determined by the value of R_2 relative to the 10kΩ of R_1. The maximum output voltage can be preset by changing the R_2 value.

Fig. 5.9 shows how the above circuit can be further developed to act as a general-purpose laboratory power supply that produces an output that is fully variable over the range 3 to 30V at currents up to 2A. Here, the reference voltage reaching the *NI* terminal of the i.c. is fixed at slightly less than 3V by potential divider R_5–R_6, and the voltage gain of the amplifier section of the i.c. is variable from unity to approximately 13:1 via the R_1–R_2 variable divider, thus enabling the regulated output voltage of the circuit to be varied from below 3V to well above 30V. The output current of the circuit is limited at 2A via R_4. Note that the circuit must be powered from an unregulated supply in the range 40V to 50V, and that the i.c. is protected against excessive supply voltages by R_8 and zener diode ZD_1.

Finally, to complete this last chapter of this volume, Fig. 5.10 shows how the 723 regulator i.c. can be used to make a practical 60V regulator that gives a protected 400mA output current. This basic circuit can be modified to give any regulated voltage up to about 250V.

In this circuit the regulator i.c. is 'floating', and the difference between the required output voltage and the unregulated supply line potential is developed across the i.c. Because of the supply voltage limitations of the i.c., the unregulated supply of the circuit must be within the range 10V to 36V above the required output voltage. Zener diode ZD_1 ensures that the voltage across the i.c. can not exceed 36V.

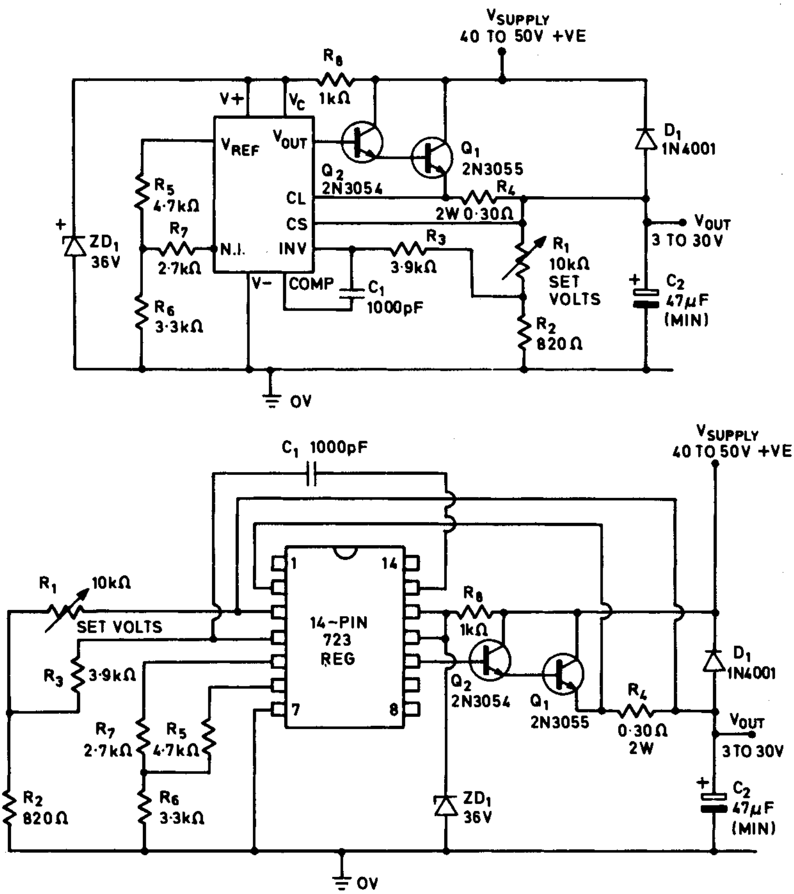
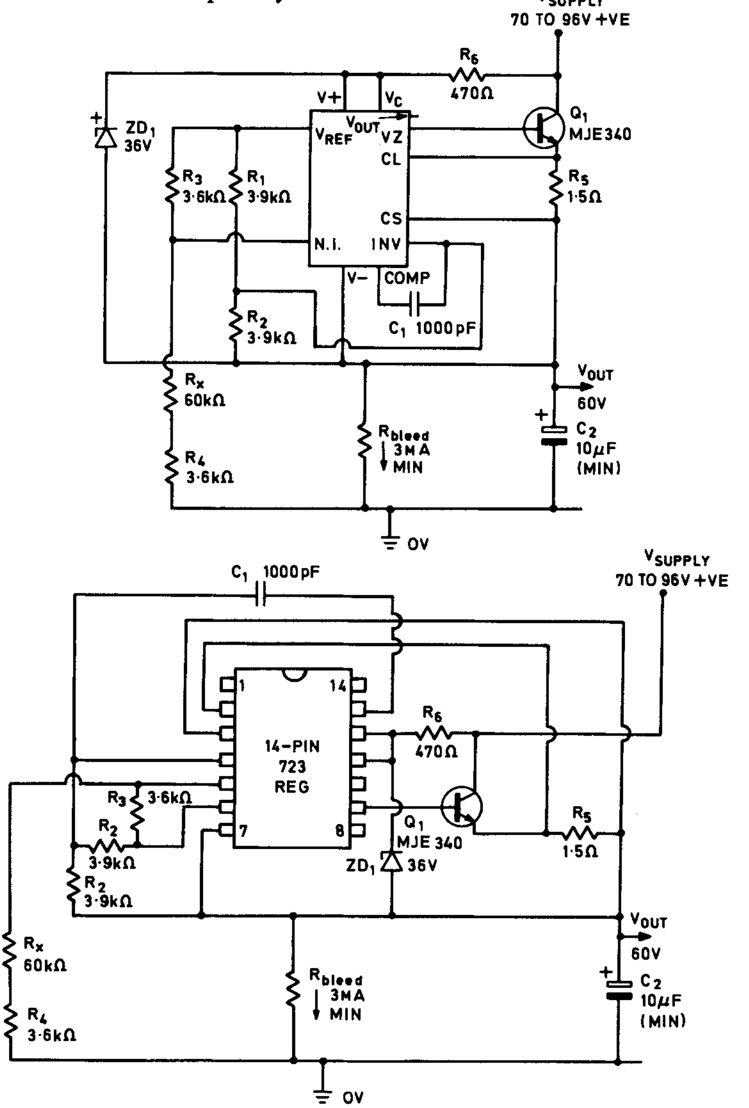


Fig. 5.9. General purpose laboratory power supply gives 3V to 30V output at currents up to 2A

The component values shown in the circuit are such that the regulated output voltage is directly proportional to the R_X value, and approximates 1V output per kΩ of value.

Note that a 'bleed' current of about 3mA passes through the i.c. when it is operating correctly, and normally this current passes through the external output load of the circuit: should this load have too high an impedance to pass this current, a separate bleed resistor (R_{bleed})

must be wired into the circuit as shown. Note that the MJE 340 output transistor used in the circuit is a 300V, 20W device that has a maximum current capability of 500mA.

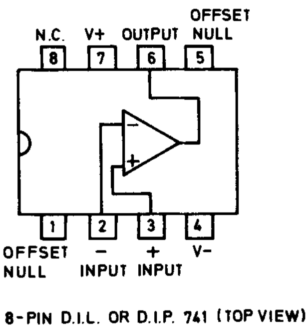


Notes: $V_{supply} = V_{out} + 10V$ to $V_{out} + 36V$
$V_{out} = V \times R_x$ (kΩ) within the limits $V_{in} - 10V$ to $V_{in} - 36V$

Fig. 5.10. 60V regulator gives protected 400mA output: circuit can be made to give outputs in the range 4V to 250V

APPENDIX

Semiconductor Outlines and Pin Designations



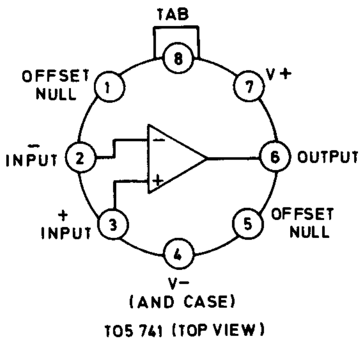


Fig. 6.1. Outlines and pin connections of the two most popular 741 op-amp packages

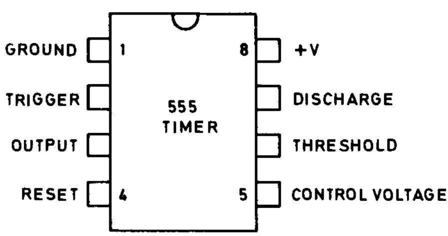


Fig. 6.2. Outline and pin notations of the standard 8-pin d.i.l. (d.i.p.) version of the 555 timer i.c. (top view)

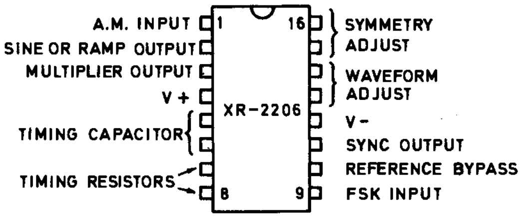


Fig. 6.3. Outline and pin notations of the 16-pin d.i.l. (d.i.p.) XR-2206 function generator i.c. (top view)

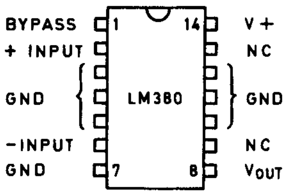


Fig. 6.4. Outline and pin notations of the standard 14-pin d.i.l. (d.i.p.) version of the LM380 i.c. (top view)

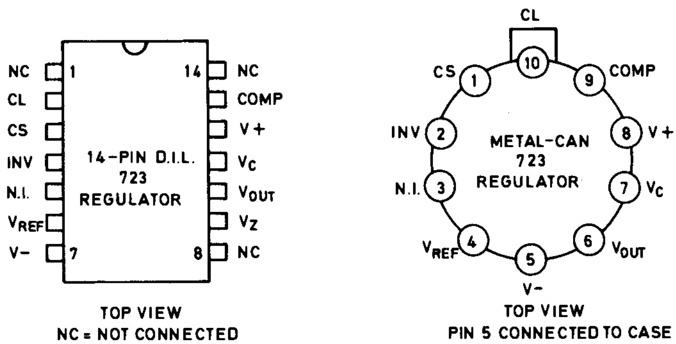


Fig. 6.5. Outlines and pin notations of the two most popular 723 regulator packages

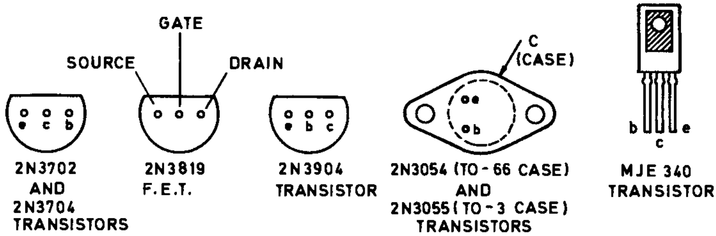


Fig. 6.6. Transistor and f.e.t. connections

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