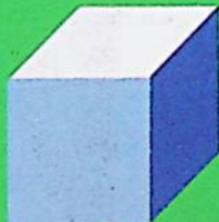
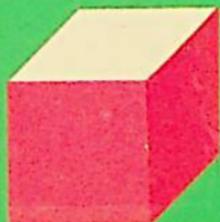
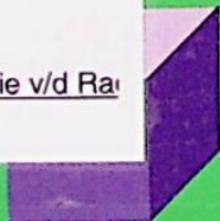
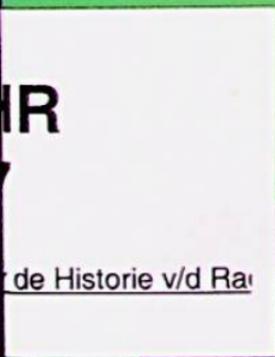
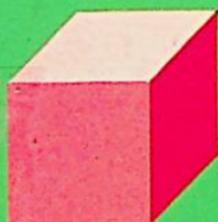
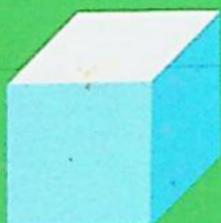
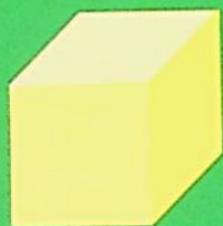


Practical Electronic Building Blocks Book 1

R. A. PENFOLD



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PRACTICAL ELECTRONIC BUILDING BLOCKS

Book 1

BIBLIOTHEEK
N.V.H.R.

by
R. A. PENFOLD

BERNARD BABANI (publishing) LTD
THE GRAMPIANS
SHEPHERDS BUSH ROAD
LONDON W6 7NF
ENGLAND

Although every care has been taken with the preparation of this book, the publishers or author will not be held responsible in any way for any errors that might occur.

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First Published – April 1983

British Library Cataloguing in Publication Data
Penfold, R. A.

Practical electronic building blocks.

Bk. 1

1. Electronic circuits

I. Title

621.381513 TK7867

ISBN 0 85934 092 9

Printed and bound in Great Britain by Cox & Wyman Ltd, Reading

Preface

BIBLIOTHEEK
N.V.H.R.

Virtually any electronic circuit will be found to consist of a number of distinct stages or *building blocks* if it is carefully analysed. This is usually quite obvious when examining a digital circuit, but is equally true (if less obvious) in relation to linear designs. Rather than gates, shift registers, and the like, linear circuits are usually comprised of filters, amplifiers, oscillators, monostables, etc.

While it would be misleading to claim that with the aid of a book giving details of numerous electronic building blocks it would be possible for the reader to design any desired linear project, a book of this type can be of great value to those who like to design and build their own projects, or to experiment with electronic circuits. Inevitably there are many projects which have one or more unusual stages which require special circuitry and components which cannot be covered in such a book, but even then there are likely to be stages which are of a standard nature and it is highly unusual for a project to need specialised circuitry throughout.

The purpose of these books then, is to provide details of a number of electronic building blocks including information such as component values and type numbers, plus details of how to change the parameters of circuits (voltage gain of amplifiers, operating frequency of oscillators, etc.) where appropriate. Some previous knowledge of electronics is assumed, and it would not really be feasible for a complete beginner to use a book of this type to full advantage anyway. However, there is no advanced theory or mathematics used, and all that is really needed is a basic understanding of the more common electronic components, together with a basic knowledge of voltage, current, impedances etc. This is mainly necessary to help the reader successfully interface one building block to another with no difficulty.

It is possible to have a great deal of fun by designing, adapting, and experimenting with electronic projects, even if the end result is not always as envisaged at the outset, and you can rapidly learn a great deal about electronics in this way.

R. A. Penfold

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Chapter 1

OSCILLATORS

Oscillators are used in a surprisingly large number of electronic projects including such diverse pieces of equipment as radio transmitters and receivers, test equipment, musical effects units, and model train controllers. It is not possible to give a design for a universal oscillator which is suitable for any application since the requirements vary greatly from one application to another. The output waveform may be irrelevant, but in most applications it is crucial for the waveform to be correct. The design of an oscillator must be altered to suit the operating frequency concerned, since a low frequency oscillator would probably be hopelessly unstable or fail to operate at all at high frequencies. Conversely, RF oscillator configurations will work at low frequencies, but require large and expensive inductors which are likely to make them a rather impractical prospect.

In this chapter a number of different oscillator circuits will be described, covering most output waveforms that are ever likely to be required, and covering a frequency range of less than 1 Hertz to many Megahertz. There are so many oscillator configurations that it is quite impossible to cover them all here, but you should find an oscillator circuit to suit all but the most specialised of applications.

555 Oscillators

The 555 timer integrated circuit has been featured in a great many projects for the electronics constructor, and this is not surprising when one considers the low cost and versatility of this device. It is well worthwhile being familiar with the capabilities of this device, and it will be considered in some detail in this book.

The standard 555 astable configuration is shown in Figure 1, and will probably be familiar to most readers. It is really a form of relaxation oscillator with C2 first charging to two thirds of the supply voltage through R1 and R2, and then discharging

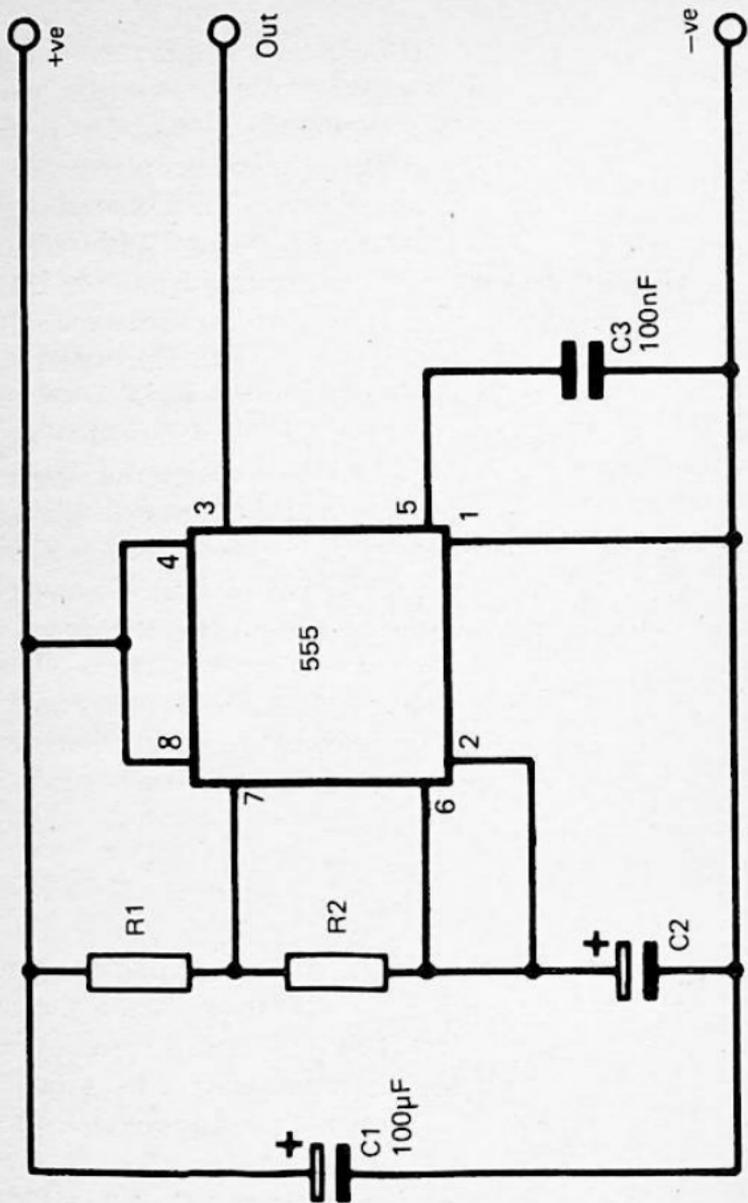


Fig. 1. The basic 555 astable configuration

through R2 to one third of the supply voltage. The operating frequency is therefore set by the values given to R1, R2 and C2, and is given by the formula:—

$$\text{Frequency} = 1.44 / (R1 + 2R2) C2 \text{ Hz}$$

The reason for adding the value of R1 to double the value of R2 (rather than simply adding the two values together) is that R2 is in both the charge and discharge paths of C2 and therefore has double the effect of R1. Rather than calculating the operating frequency with the resistances in ohms and the timing capacitance in farads it is better to use Megohms and microfarads. This avoids having very low figures for the capacitance and very large figures for the resistance.

Typical values for the timing components would be 4k7 for R1 and R2 plus 100nF for C2 giving an operating frequency of just over 1kHz (4k7 plus twice 4k7 = 14k1 or 0.0141M, multiplied by 100nF or 0.1μF = 0.00141, and 1.44 divided by 0.00141 = 1021Hz or 1.021kHz. The frequency of operation can be changed by altering the value of C2, with changes in value giving an inversely proportional change in frequency (e.g. doubling the value of C2 halves the operating frequency). Similarly, the operating frequency can be changed by altering the timing resistance with changes again giving an inversely proportional variation of the frequency of the circuit.

There are, of course, certain limitations on the circuit values, and it is advisable to keep the combined resistance of R1 and R2 to no more than about 10 Megohms. A resistance substantially higher than this could give rather unpredictable results or might result in the circuit failing to oscillate at all. This is due to the very low currents that would be present in the timing circuit, and consequent effects tiny input currents drawn by the 555 or leakage through C2 might have.

There is also a limit on the minimum value R1 should have, and this can be calculated by dividing the supply voltage by 0.2. This would, for example, give a figure of 45 ohms with a 9 volt supply. In practice it is often necessary to use a value very much higher than this absolute minimum figure, and one reason for this is simply to keep the dissipation in the 555 down to a reasonable level. It should also be remembered that an internal

transistor of the device places a virtual short circuit from pin 7 of the 555 to the negative supply rail during the discharge period of C2, and apart from any effects on the 555 itself, a low value component in the R1 position will give a high dissipation in this resistor and a very high current consumption during the discharge period. A minimum resistance of about one kilohm is therefore more realistic.

Pin 3 is the output of the 555, and this terminal goes high while C2 is charging and low when it is discharging. This gives a rectangular waveform at the output of the device, but it is not a squarewave and this basic configuration cannot produce a proper squarewave having a one-to-one mark-space ratio. This is simply because C2 charges through both R1 and R2, but it discharges through R2 alone (plus the insignificant impedance of the discharge transistor within the 555). The charge (high) period must therefore be longer than the discharge (low) one. In some applications this may be desirable, and in many others it may be irrelevant, but as we shall see later it is possible to obtain any desired mark-space ratio using a suitably modified configuration.

The output of the 555 gives a peak to peak voltage almost equal to the supply voltage used, and it is at a low impedance. The device has a form of class B output stage, and it can source or sink up to 200mA. However, with high output currents the output is pulled well away from its normal (no load) level, and a combination of large output voltage and high current cannot be supplied directly by the device. This is easily overcome by the addition of an emitter follower buffer stage at the output though.

An output signal can be taken from across C2, but this is only permissible if a very low output current is taken since any loading here obviously has an effect on the charging and discharging of C2. The higher the values of R1 and R2, the less output current that is available. The waveform across C2 is roughly triangular, but C2 charges and discharges exponentially and so this is obviously not a linear triangular waveform. Also, like the output at pin 3, the mark-space ratio of the signal is not one-to-one. Despite these shortcomings, the signal across C2 is often useful in applications such as modulated tone generators

and musical effects units.

A slight shortcoming of the 555 is that once per output cycle the supply is "crowbarred" and a negative voltage spike is introduced onto the supply lines. C1 is included to decouple this noise, although in some applications there may be no stages in the project that will be adversely affected by these noise spikes, and C1 would not then be necessary. Note that even with C1 included in the circuit the noise spikes may not be totally suppressed.

C3 is also a decoupling capacitor, and this removes any stray pick-up at pin 5 of the 555. A control voltage applied to pin 5 can be used to frequency modulate the oscillator, as we shall see later, and the decoupling is needed to avoid unwanted frequency modulation due to stray pick-up. Unless the device is in a position where it is likely to pick up a lot of strong noise it is unlikely that C3 will have any noticeable effect on the circuit though, and it is usually omitted.

Any electronic circuit will only operate between certain supply voltage limits, and this is something which must always be borne in mind. For the 555 device the minimum and maximum supply potentials are 5 and 15 volts. The 555 actually has a maximum permissible supply voltage of 16 volts, but with any figure of this type it is advisable to leave a small safety margin and 15 volts is therefore the maximum nominal supply voltage. The maximum operating frequency of the device is at least 500kHz, and there is no properly defined lower limit to the operating frequency. The lower limit is determined by the quality of the timing capacitor C2 which for very low frequencies must be a polarised type such as a tantalum or electrolytic type. Although in theory it is perfectly alright to use a capacitor of (say) 100 μ F with a total timing resistance of several megohms, in practice this is unlikely to work. The leakage resistance of the capacitor may well be only a few hundred kilohms, and this would limit the charge voltage to a level which would not permit correct operation of the circuit. Tantalum types are generally much better in this respect than electrolytic capacitors, and such a combination of timing component values might work well using a tantalum capacitor. However, as the highest value tantalum capacitors that are generally available are

not above $100\mu\text{F}$ or so in value, a substantial increase in the timing capacitors value is not feasible unless the timing resistance is reduced, which would counteract the higher capacitance value. This limits the minimum operating frequency to about one cycle every ten minutes or thereabouts, but in practice it is unlikely that a lower operating frequency would ever be required.

In low frequency applications it is often more convenient to work in terms of the output high and low times rather than in terms of frequency. The high time is equal to:-

$$T_h = 0.685 (R_1 + R_2) C_2$$

The low time is equal to:-

$$T_l = 0.685 R_2 C_2$$

There is no minimum limit on the value of C_2 incidentally, and with no discrete capacitor here the circuit will almost certainly oscillate at a frequency determined by the internal capacitance of the 555 of around 30pF . In order to obtain reliable and consistent results it is advisable to use a timing capacitor of no less than about 1nF in value.

A point that is sometimes of importance, especially when the 555 is used in very low frequency applications, is that the first half of the initial cycle is longer than that obtained on subsequent cycles. This is simply due to C_2 starting with zero charge on the first cycle, whereas it starts with a charge equal to one third of the supply voltage on subsequent cycles. This is not unique to the 555 by any means: in fact most simple C – R oscillators produce a longer initial cycle.

Gated Oscillator

The 555 can be used as a gated oscillator, as shown in Figure 2. This circuit is basically the same as the one shown in Figure 1, but instead of pin 4 of the 555 being taken to the positive supply rail it is fed with the gating signal. Pin 4 must be taken below about 0.5 volts in order to switch off the oscillator, and the output goes to the low state when the circuit is gated off. The input current to pin 4 is only about $100\mu\text{A}$. If pin 4 is

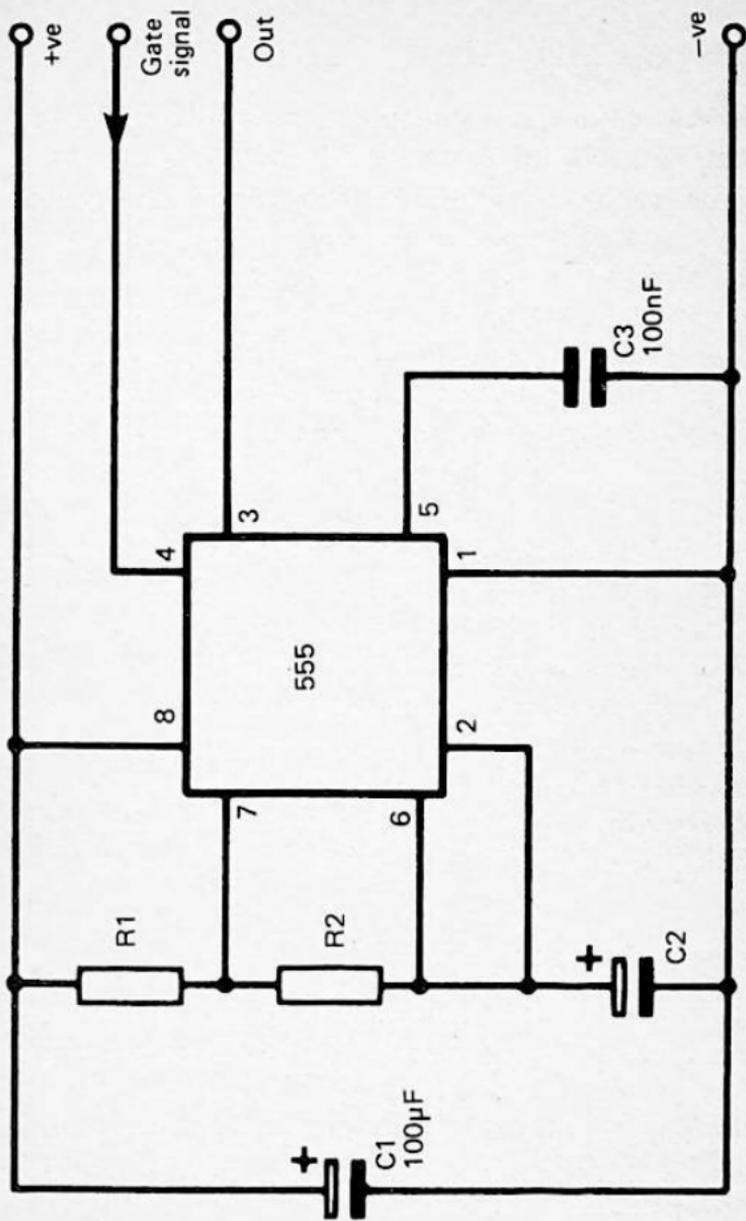


Fig. 2. A gated 555 astable circuit

left floating it is normal for the oscillator to operate, incidentally.

The current consumption of the 555 varies from about 3mA with a 5 volt supply to about 10mA with a 15 volt supply (it is about 7 to 8mA with a 9 volt supply).

Adjustable Duty Cycle

As mentioned earlier, it is possible to produce a 555 astable circuit that can give any desired mark-space ratio, and a circuit of this type is shown in Figure 3. This has steering diodes D1 and D2 which direct the charge current through R1 and the discharge current through R2. Thus a 1 : 1 mark-space ratio can be obtained by simply making R1 and R2 the same value. If R1 is made larger in value than R2 the charge (high) time is made proportionately longer than the discharge (low) time. Similarly, making R2 higher in value than R1 makes the discharge (low) time proportionately longer than the charge (high) time.

The formulas for calculating the operating frequency and high output time are slightly different since R2 is not in circuit while C2 is charging. Thus in the frequency calculation the timing resistance is simply R1 plus R2 rather than R1 plus double R2. Similarly, the timing resistance when calculating the high output time is simply R2 and not R1 plus R2. The two diodes have a marginal effect on the frequency, charge, and discharge times, but this will not normally be large enough to be of significance.

Frequency Modulation

As mentioned earlier, the operating frequency of a 555 astable can be modulated by a control voltage, and the circuit for this is shown in Figure 4. This differs from the normal 555 astable configuration only in that the decoupling capacitor at pin 5 of the device has been removed and the control voltage is applied to this pin instead.

Pin 5 of the device connects to the internal potential divider circuit which sets the two thirds of the supply voltage threshold level at which the 555 triggers from the charge mode to the

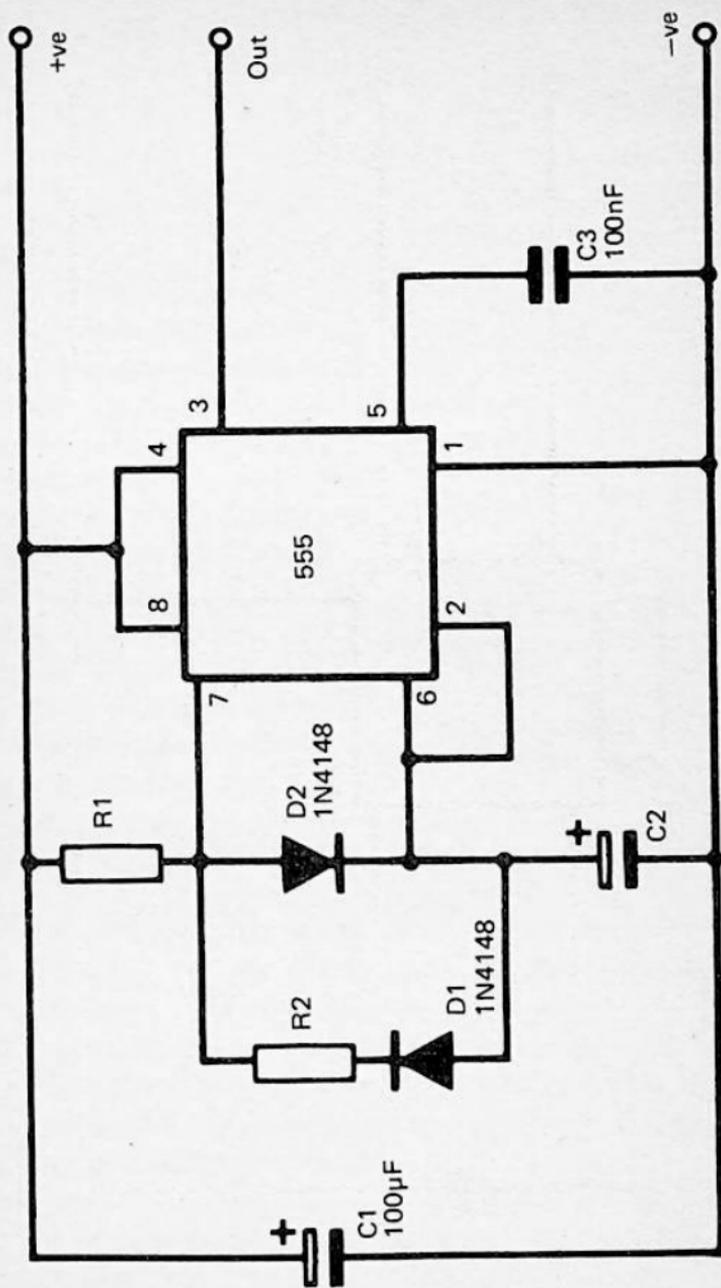


Fig. 3. An adjustable mark-space ratio 555 astable

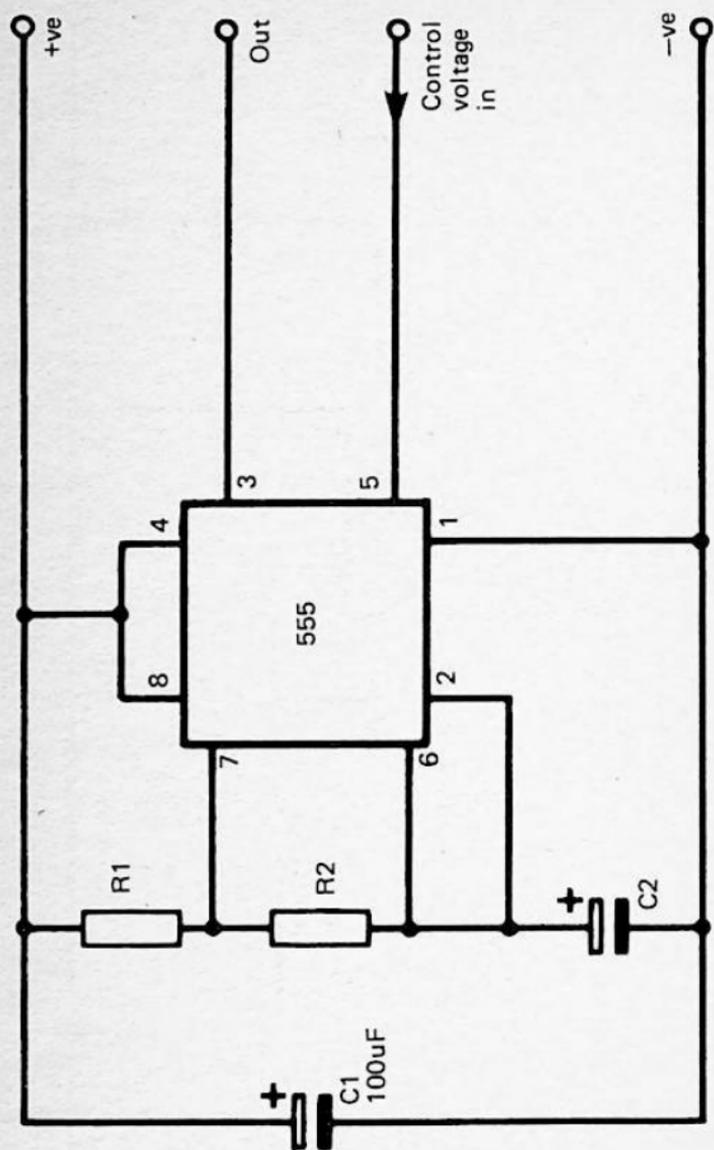


Fig. 4. A simple 555 V.C.O. circuit

discharge one. By taking pin 5 above two thirds of the supply voltage the operating frequency is decreased since C2 has to charge to and discharge from a greater potential, and therefore takes longer to do so. Taking pin 5 below two thirds of the supply voltage reduces the voltage to which C2 has to charge and discharge so that both processes take less time and the operating frequency is increased. The output waveform at pin 3 remains unchanged, but there is a slight change in the wave-shape and the amplitude of the signal across C2.

Pin 5 of the 555 does not connect directly to the internal potential divider of the device, and the threshold voltage at which the device triggers from the charge state to the discharge one is not equal to the voltage fed to pin 5. This limits the degree of control that can be obtained using a control voltage fed to pin 5, and the frequency is reduced by a little more than 50% if the control input is taken to the positive supply voltage. The output frequency is slightly more than doubled if the control terminal is taken to one third of the supply voltage, and it can be boosted a little further if it is taken slightly lower in voltage. However, if it is taken too low in voltage a malfunction occurs and the output frequency reduces considerably. It is therefore advisable to keep the control voltage between one third of the supply voltage and the full supply potential.

The control range of this circuit is not very great but it is useful for applications such as modulated tone generators.

Tone Generator

The 555 is a useful device for applications where an audio tone of moderate volume is required, and it is possible to drive a loudspeaker direct from the output of the device. This can be rather inefficient though, and if a low impedance loudspeaker is used it can even result in a malfunction of the circuit.

It is normally better to use capacitive coupling at the output, as shown in Figure 5. Although a high impedance loudspeaker is specified in Figure 5, the circuit will in fact work with a low impedance 8 ohm type although for most applications a high impedance speaker will give adequate volume. An output power of a few hundred milliwatts will be produced into an 8

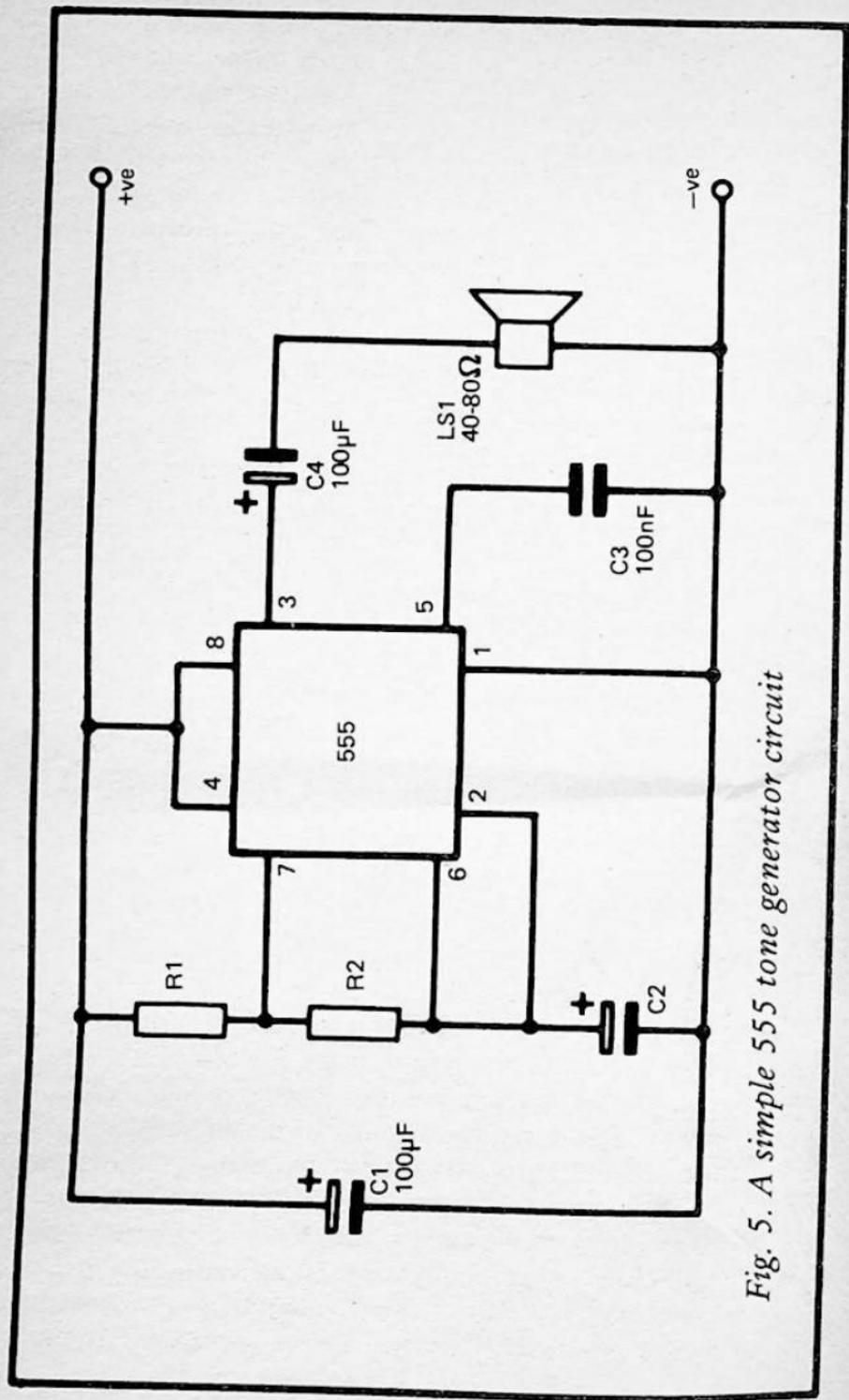


Fig. 5. A simple 555 tone generator circuit

ohm load, and care has to be taken here since many miniature loudspeakers can only handle about 200mW or so.

Of course, you are not restricted to simply using the 555 as a gated astable, or as a voltage controlled oscillator, and you can combine the circuits described so far to give (say) a gated, frequency modulated tone generator, or a gated squarewave generator. It is this versatility plus its low cost that has made the 555 such a popular device.

CMOS 555

There is a very useful version of the 555 timer IC known as the ICM7555, or it is sometimes just called the 7555. It differs from the standard 555 mainly in that it has CMOS circuitry which gives the device a low current consumption of only about 80 microamps, and the inputs of the device require far lower input currents. The lower current consumption is obviously useful for battery powered equipment where the higher current consumption of the standard 555 can give quite high running costs, especially with circuits that employ more than one 555. However, bear in mind that, as with the standard 555, there is current consumed by the timing network in addition to that taken by the device itself, and this current must obviously be kept low (by using high timing resistor values) in order to take full advantage of the device's low current consumption. Of course, any output current that flows also boosts the current consumption of the circuit, and may make the reduced current drain of the 7555 largely irrelevant. The 7555 is substantially more expensive than the standard 555 and it is pointless to use the 7555 except where it will give genuine and worthwhile advantages.

The advantage of having low input currents to the 7555 is primarily that it enables very high timing resistances to be used if desired, and values as high as 100 Megohms are quite acceptable. It also means that when used as a gated oscillator pin 4 can be fed from a high impedance source if necessary since the input current here is totally insignificant.

Another advantage of the 7555 is its lower internal capacitance which enables a timing capacitance of as little as 100pF to be

used with consistent and reliable results. It also has a wider supply voltage range of 2 volts to an absolute maximum of 18 volts.

The main disadvantage of the 7555 is that it has reduced output current capability in comparison to the standard 555. The sink current at around 100mA maximum is still quite high and is likely to be more than adequate for applications where the load is connected between the output and the positive supply rail. The situation is different for loads connected between the output terminal and the negative supply rail since the source current of the device is only a few milliamps. Obviously the 7555 cannot be regarded as a direct substitute for the standard 555 because of this, and it may sometimes require a discrete buffer stage at the output where the 555 would not. The ICM7555 can be used with the tone generator circuit of Figure 5, but it is not advisable to use a low impedance loud-speaker, and the volume obtained will be appreciably lower than that obtained using an ordinary 555.

Despite the fact that the 7555 is a CMOS device it does not require the handling precautions normally associated with CMOS integrated circuits as this device has fully effective protection circuitry built-in.

One final point about the 7555 is that it does not "crowbar" the power supply to anything like the extent that the normal 555 does, and it therefore needs little or no supply decoupling.

Op. Amp. Oscillator

Figure 6 shows the circuit diagram of a simple squarewave oscillator which is based on a single operational amplifier. The specified values give an operating frequency of approximately 1kHz, but this can be altered by changing the value of R1 or C1, with changes in value giving an inversely proportional alteration in the output frequency. The following equation enables the approximate output frequency to be calculated:-

$$\text{Frequency} = 0.82/R1C1 \text{ Hz}$$

R1 can have any value from a few kilohms to many Megohms, and C1 can have a capacitance value of anything from a few

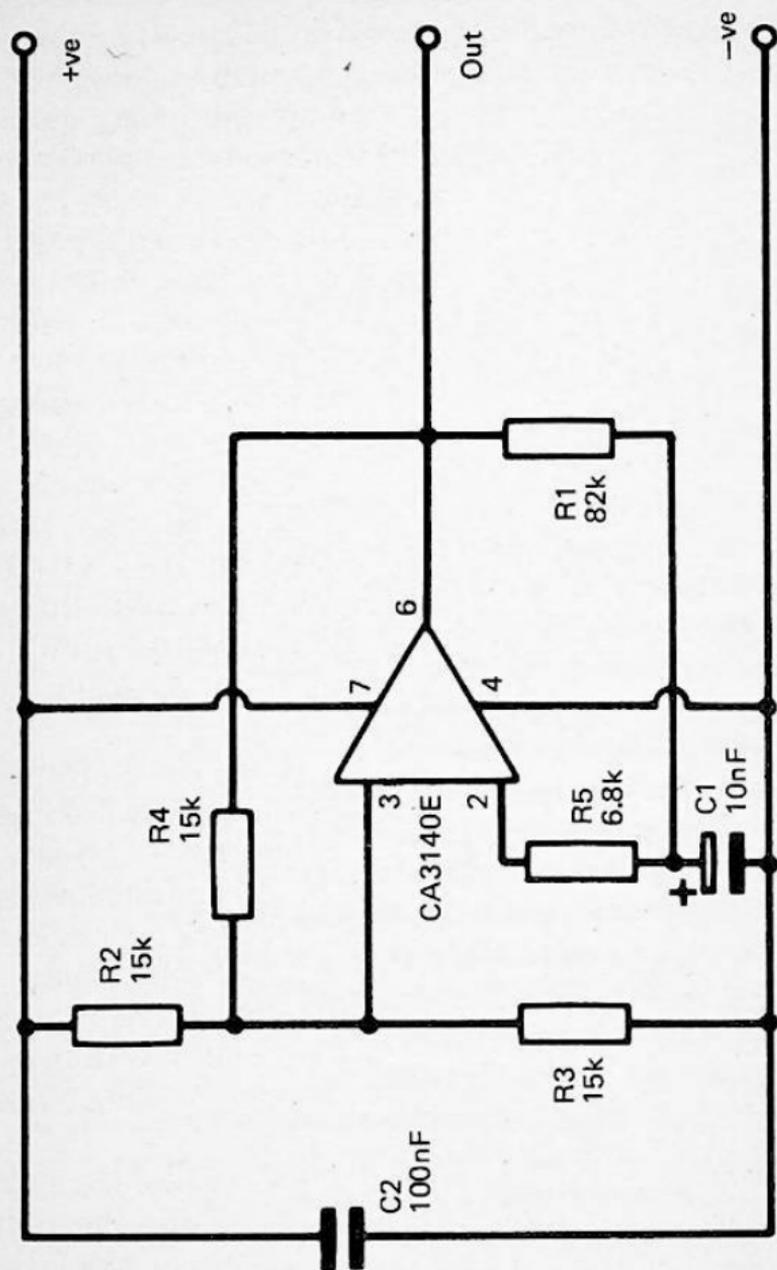


Fig. 6. The circuit of a simple op. amp. squarewave oscillator

tens of pF upwards. The diagram shows the correct polarity for C1 if a polarised type such as an electrolytic or a tantalum component is used here.

The maximum operating frequency is 100kHz or more, but at high frequencies the output waveform tends to deteriorate somewhat due to limitations placed on the circuit by the slew-rate of the CA3140E. An operational amplifier does not have a high current output stage, and without additional amplification at the output of the circuit it is possible to take a maximum output current of only a few milliamps.

Like 555 astables, this circuit is quite stable with regard to output frequency, and changes in supply voltage do not produce significant frequency shifts. The circuit will operate well using a supply potential down to about 6 volts, and the absolute maximum supply voltage is 36 volts (which is the maximum figure for the CA3140E operational amplifier).

The circuit can employ a different operational amplifier, such as an LF351, TL081CP, etc., but note that many of these give a lower peak to peak output voltage than the CA3140E (which gives an output voltage swing almost equal to the supply potential under low load conditions), and a side effect of this is a marginal increase in the output frequency of typically around 10%.

This circuit is useful where a variable frequency squarewave oscillator is required since the operating frequency can be varied by a single resistance, and it is merely necessary to replace R1 with a fixed resistor and a variable resistor connected in series. Of course, the 555 astable circuits described earlier can also be used as variable frequency oscillators if one of the timing resistances is replaced by a fixed resistor and a variable resistor in series, but this gives a change in the output waveform as the output frequency is adjusted, and this might be undesirable in some applications.

In common with the 555 astables described earlier, a non-linear triangular waveform is available from across the timing capacitor. Again this is at a high impedance, and very little output current is available.

Pulse Generator

The circuit of Figure 6 can be easily modified to produce a pulse output waveform, and Figure 7 shows the necessary modifications. This really just consists of using two timing resistors (R1 and R2) plus steering diodes D1 and D2 so that C1 charges via R2 and discharges via R1. Thus R1 sets the low output period and R2 sets the high output period, and the mark-space ratio is equal to the ratio of R1 to R2.

The operating frequency of this oscillator can be calculated using the formula given for the oscillator circuit of Figure 6, but the timing resistance is the average of R1 and R2 in this case.

Sinewave Generator

Although the sinewave generator circuit of Figure 8 is extremely simple it nevertheless provides an extremely pure output which has a typical total noise and distortion level which is well below 0.1%. It is a straight forward Wien Bridge oscillator which is based on an operational amplifier and has thermistor Th1 to stabilise the closed loop gain of the circuit at a level which gives a high quality output signal having an amplitude of approximately 2 volts peak to peak. The only real disadvantage of this type of circuit is that the RA53 thermistor (also known as just an R53) is a special self heating type which is far more expensive than most other thermistors. However, the simplicity of circuitry it permits plus the very high quality output that can be obtained using this method of stabilisation probably justifies the expense in applications where a very high quality sinewave is required.

C1, C2, R1 and R2 are the components which determine the operating frequency of the circuit, and it is normal for R1 to have the same value as R2, and for C1 and C2 to be of the same value. The frequency of oscillation can be calculated using the following formula:-

$$\text{Frequency} = 1/2\pi CR \text{ Hz}$$

Thus for an operating frequency of about 1kHz C1 and C2 could be 4n7 with R1 and R2 at 33k. Changes in the value of

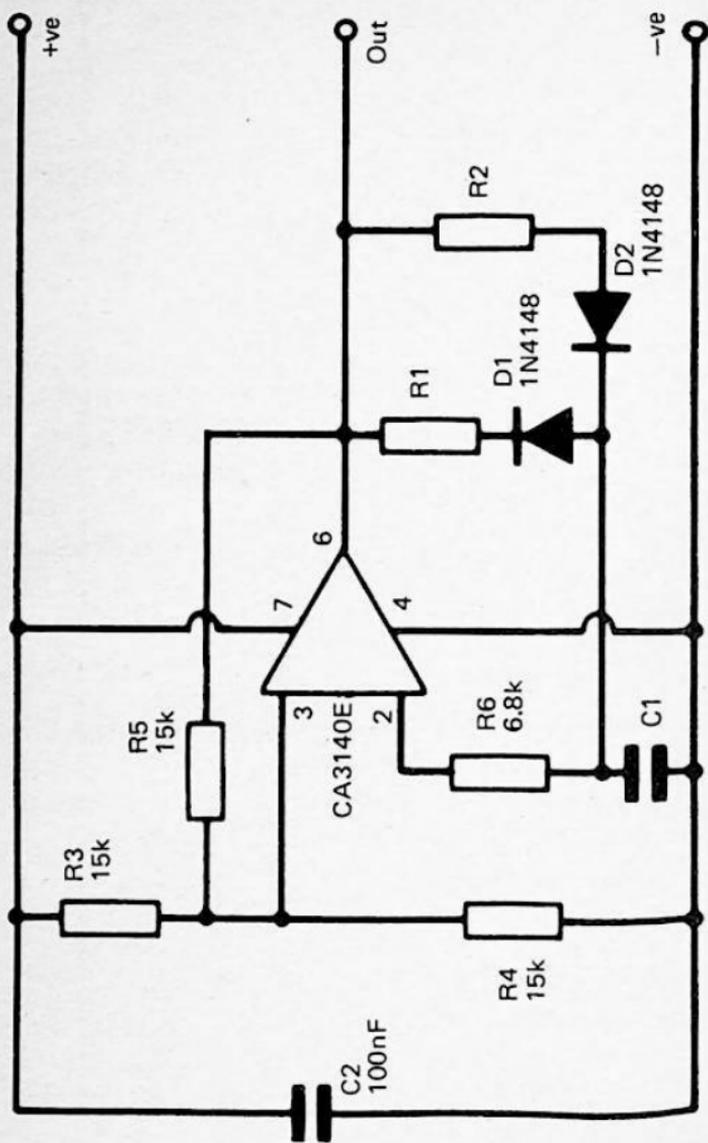


Fig. 7. An op. amp. pulse generator circuit

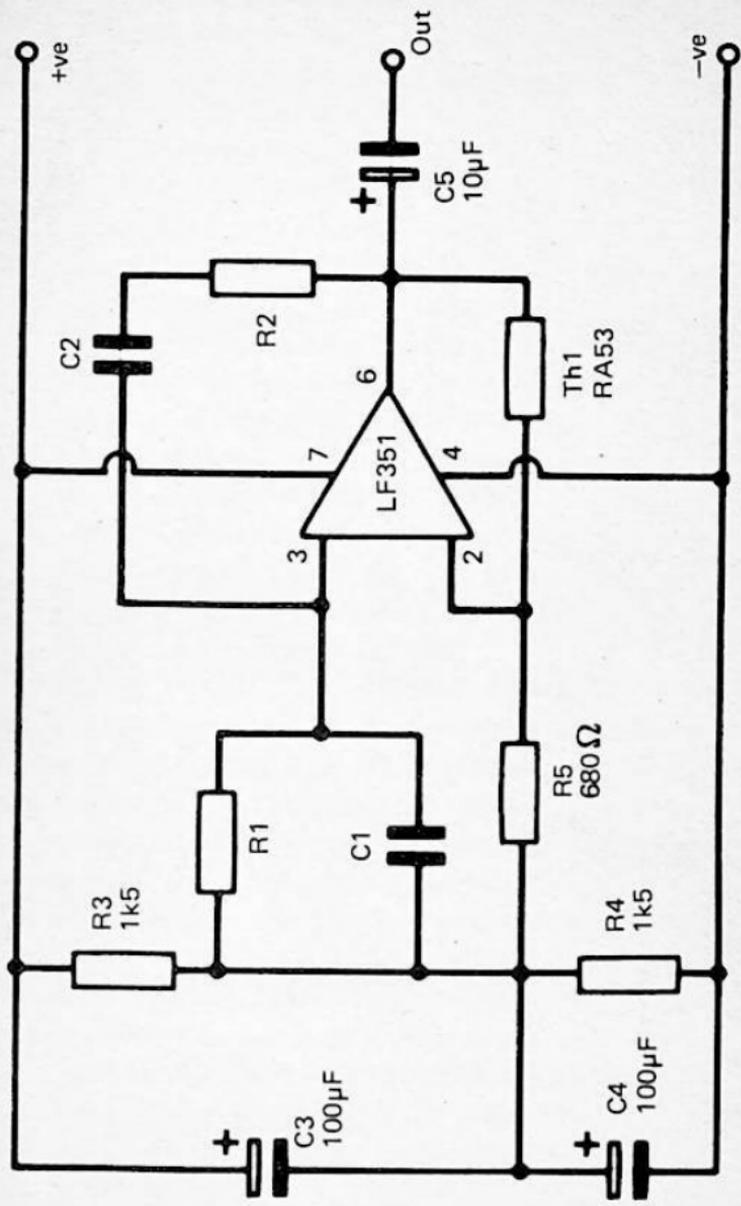


Fig. 8. A high quality sine wave generator circuit

either the resistors or the capacitors gives an inversely proportional change in frequency. The two resistors should be between a few kilohms and several Megohms in value, and the capacitors can have any value of a few picofarads or more. However, the capacitors cannot be polarised types such as electrolytic or tantalum components, and in practice this limits the maximum value to about $2\mu\text{F}$.

The operating frequency of the circuit can be made variable if R1 and R2 are replaced by a fixed resistor and a variable resistor in series, and it is of course necessary to use a dual gang potentiometer so that R1 and R2 can be adjusted together. A circuit of this type is in fact used as the basis of many high quality audio signal generator designs.

The circuit requires a minimum supply voltage of about 6 volts, and the absolute maximum is 36 volts. The circuit is effectively powered from a dual balanced power supply with the central 0V supply being produced by R3, and R4. It can be powered from proper dual supplies, of course, and R3 plus R4 then become superfluous and can be omitted.

Simple Sinewave Generator

Figure 9 shows the circuit diagram of a sinewave generator which is basically the same as that of Figure 8, but it uses a method of gain stabilisation that does not require an expensive thermistor. D1 and D2 are used to reduce the closed loop gain of the amplifier when the output voltage exceeds about plus or minus 0.5 volts, and this prevents the circuit from oscillating violently so that a heavily clipped and distorted output is avoided.

There is a significant amount of distortion on the output of the oscillator though, and it is not suitable for applications where a high quality sinewave is required. The output level of the circuit is about 500mV RMS.

Phase Shift Oscillator

A phase shift oscillator is a simple circuit which gives a sinewave output signal, but unless some form of gain stabilisation is used

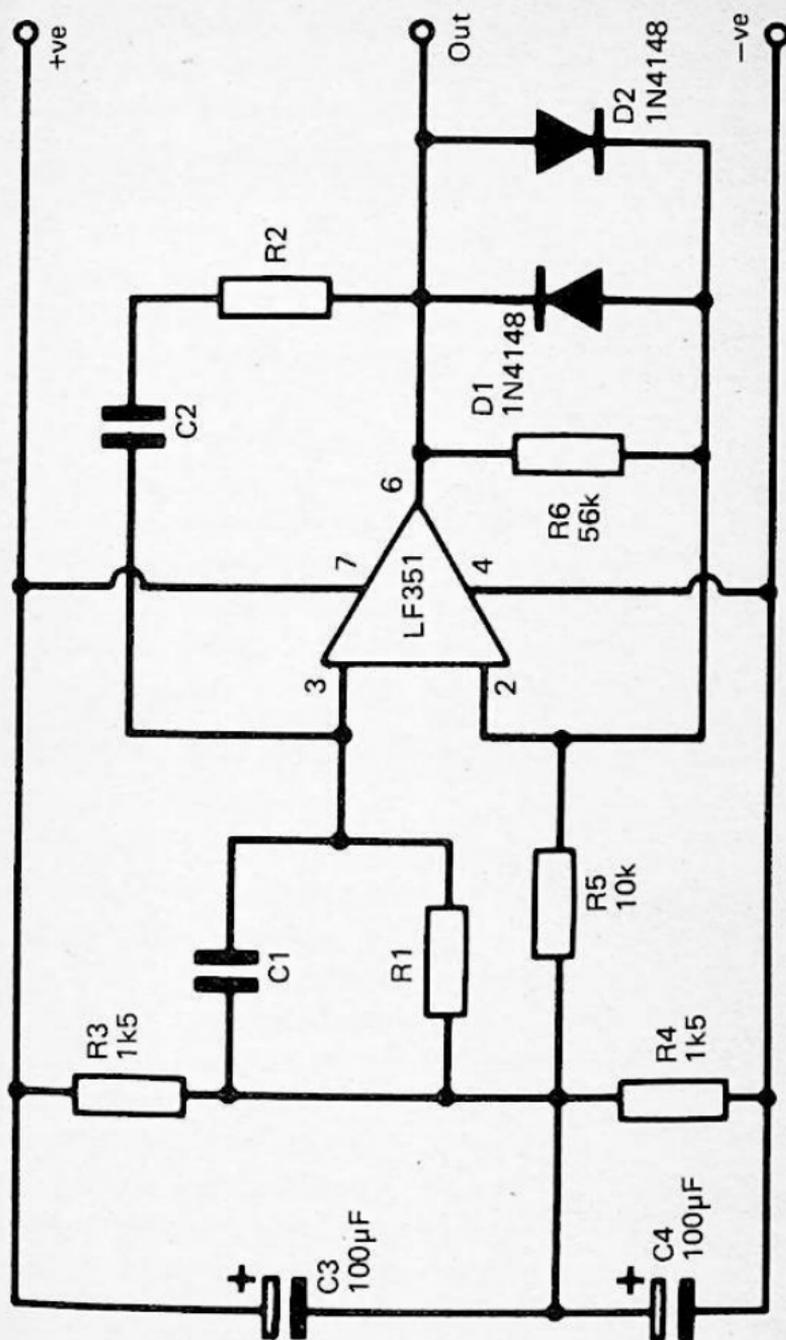


Fig. 9. A simple sine wave oscillator circuit

it gives an output that contains distortion products of significant proportions. The operating frequency of this type of circuit is controlled by three resistors and three capacitors, and it is therefore not really practical to use this type of oscillator in applications where a variable frequency facility is required. It is not easy to achieve gain stabilisation with this type of circuit either. It is therefore best to use a phase shift oscillator in applications where a fixed frequency, low quality sinewave signal is required.

The circuit diagram of a simple phase shift oscillator is provided in Figure 10. Tr1 is used as a high gain common emitter amplifier with feedback provided between the collector and base terminals by a three stage phase shift network. The first stage consists of C1 and R1, the next is C2 and R2, and the last is comprised of C3 and the input impedance of Tr1. At a certain frequency there is a phase shift of 60 degrees through each section of the phase shift network, giving a total shift of 180 degrees. Thus positive rather than negative feedback is provided over the amplifier, and the gain of Tr1 is sufficient to ensure strong oscillation at this frequency.

In theory the frequency of operation is given by the formula:-

$$\text{Frequency} = 1/2\pi \sqrt{6CR} \text{ Hz}$$

This assumes that the gain of the amplifier is just sufficient to sustain oscillation, but in practice it is substantially higher than this. The result is that the actual frequency of oscillation is much lower than the calculated figure, and the operating frequency is in fact typically only about half the calculated figure. Thus the circuit values shown in Figure 10 give an output frequency of about 1kHz instead of 2kHz.

The frequency of this circuit can be altered by changing the value of C1 to C3, and these should all have the same value. Changes in value give an inversely proportional shift in frequency. The frequency can also be altered by changing the resistor values, but it is advisable to keep R1 and R2 at no less than about 3k3, and no more than about 18k. Tr1 can be any high gain NPN silicon transistor such as a BC109C or BC108C.

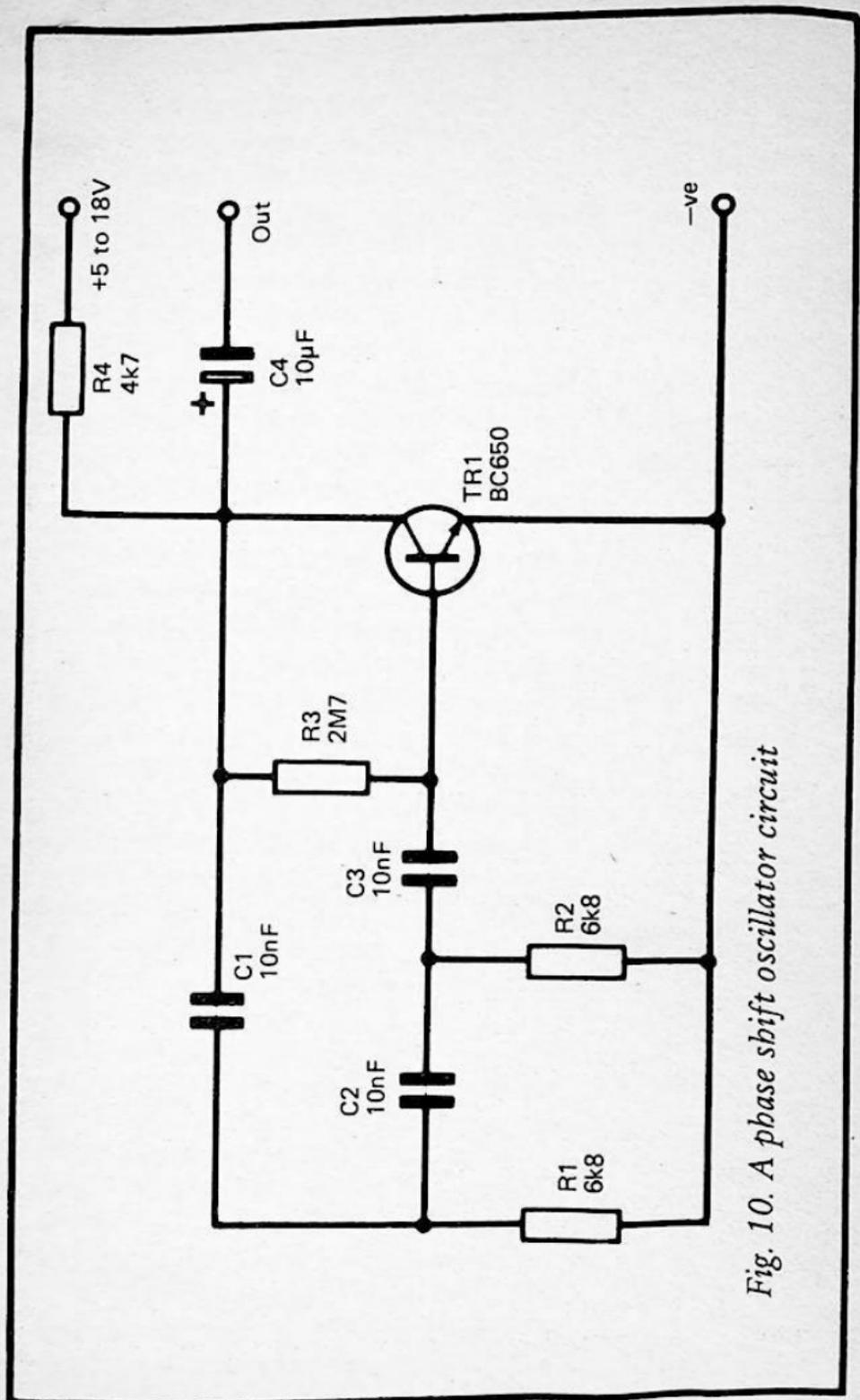


Fig. 10. A phase shift oscillator circuit

CMOS Oscillator

As we shall see later, there are CMOS logic devices which are designed specifically as oscillators, but it is possible to use certain simple CMOS logic elements such as gates and inverters as oscillators, and this is often cheaper than using one of the special devices.

Figure 11 shows how a couple of CMOS inverters can be connected to give a roughly squarewave output (the mark-space ratio will not be exactly one-to-one, but it will be reasonably accurate). R1 simply biases the inverters as linear amplifiers, and C1 provides positive feedback which produces strong oscillation and consequently an almost squarewave output.

The specified values give an operating frequency of about 1kHz, but this will vary somewhat from one device to another. Output frequency is also affected to a small but significant extent by changes in supply voltage. It is possible to modify the circuit slightly to give better stability, but in applications where highly predictable results and good stability are essential it is probably better to use an alternative type of oscillator.

The operating frequency can be altered by changing the value of C1 or R1, and in both cases the shift in frequency is inversely proportional to changes in value. C1 can have a value of anything from a few picofarads upwards, and R1 can have a value of anywhere between a few kilohms and many Megohms

The circuit can employ any normal CMOS inverters such as those contained within the 4069 device, or it is possible to use NOR or NAND gates (such as those in the 4001 and 4011 devices) if the inputs of each gate are connected together so that it gives a simple inverter function. Supply connections and other leadout numbers are not given in the circuit diagram as these will obviously depend upon the device used. The supply voltage can be anything from 3 to 15 volts, and the supply current is just a few milliamps. The circuit will operate well at frequencies as high as several hundred kilohertz, but the output waveform deteriorates at high operating frequencies, and the current consumption becomes somewhat higher.

If antiphase signals are required either an additional inverter can be used at the output to give the second, out-of-phase output

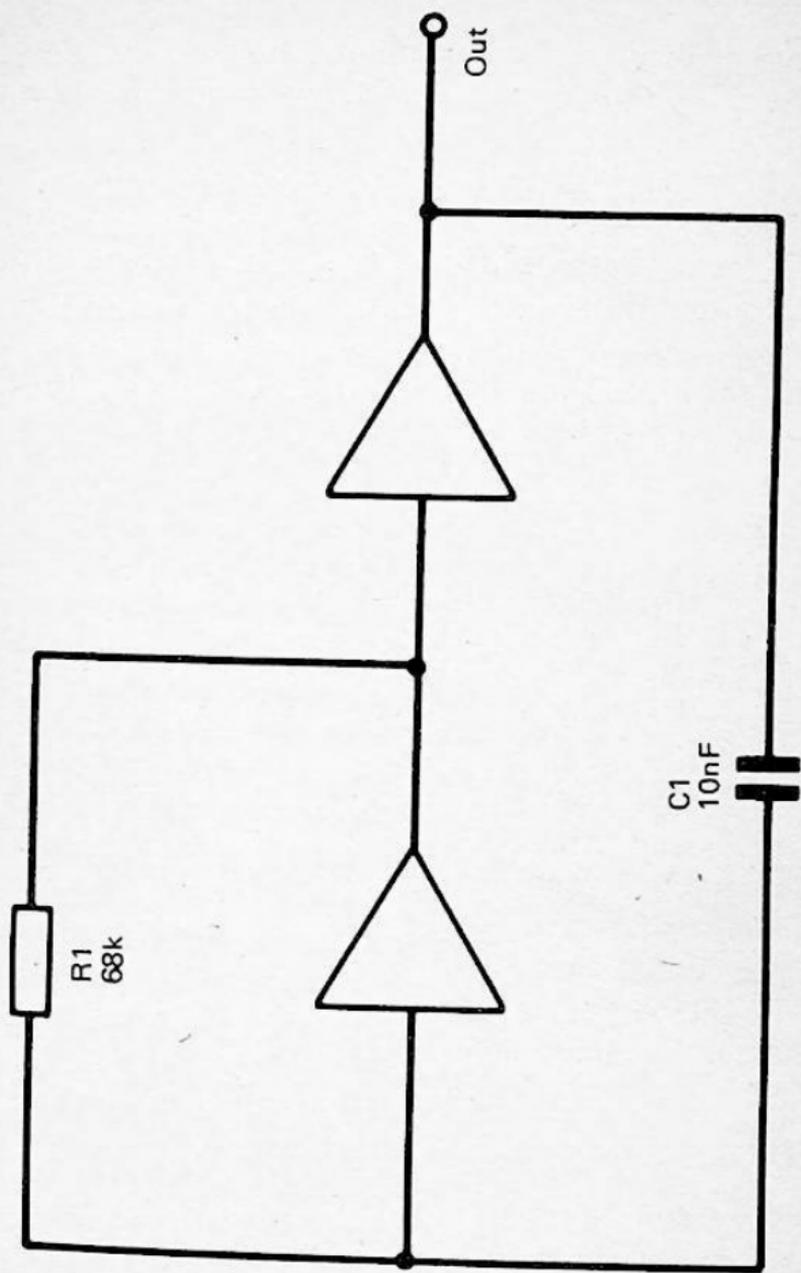


Fig. 11. A simple CMOS astable multivibrator

or a reasonably good antiphase squarewave signal can be obtained from the output of the first inverter of the astable.

Pulse Generator

The circuit of Figure 11 can be modified to act as a pulse generator, and this modification is shown in the circuit diagram of Figure 12.

Here the timing resistor has been replaced by two resistors plus steering diodes so that one resistor is operational while the output of the oscillator is high, and the other is operational while the output of the circuit is low. The low output period is governed by R_1 , and the high output period is controlled by R_2 . Thus the mark-space ratio of the output signal is roughly equal to the ratio of R_1 to R_2 .

Simple VCO

By adding a resistor at the input of the basic astable circuit of Figure 11 (as shown in Figure 13), a crude but useful voltage controlled oscillator can be produced. R_2 should have a value of about ten times R_1 , and taking the control input high then causes the output frequency to decrease somewhat while taking this input low gives an upward frequency shift. The maximum output frequency obtained is only about two or three times the minimum output frequency, but this circuit can be used in non-critical applications such as frequency modulated tone generators. Variations in the control voltage cause slight changes in the output waveform, but in most applications this will not be of any importance.

Gated Oscillators

If a CMOS astable is based on NAND or NOR gates rather than true inverters it is possible to readily produce a gated astable, and Figure 14 shows one way of achieving this.

The circuit is based on two of the NOR gates in a 4001 quad 2 input NOR gate. The first gate has its inputs connected together so that it provides an inverter action, but the second

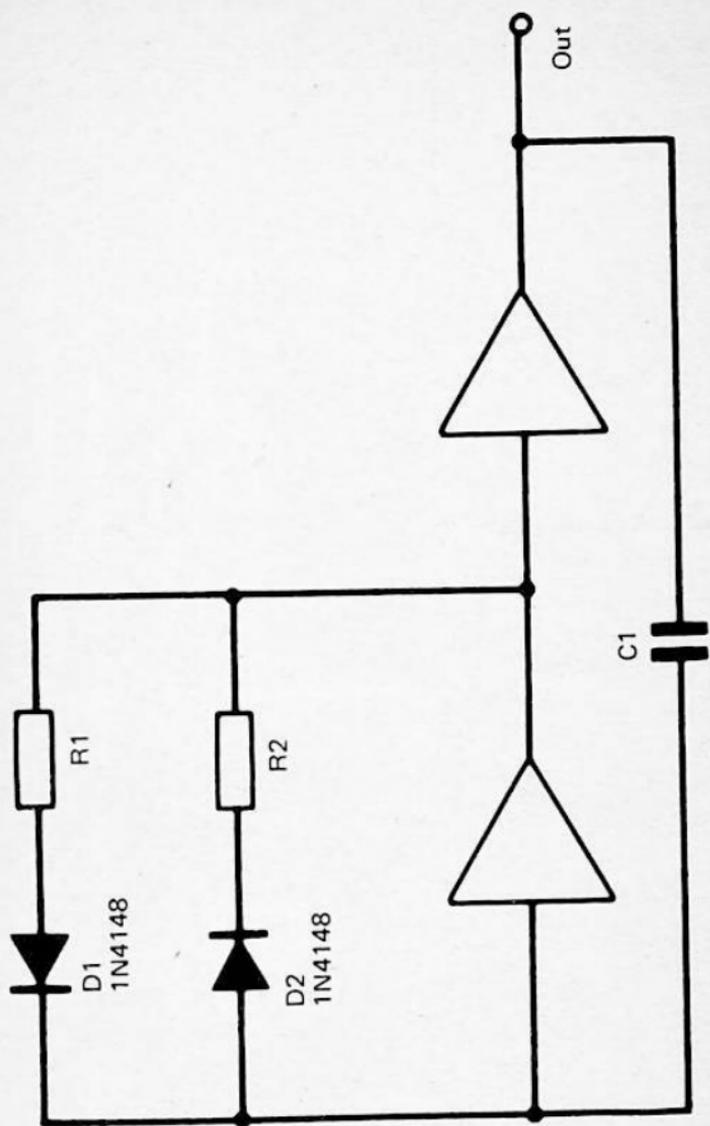


Fig. 12. A simple CMOS pulse generator circuit

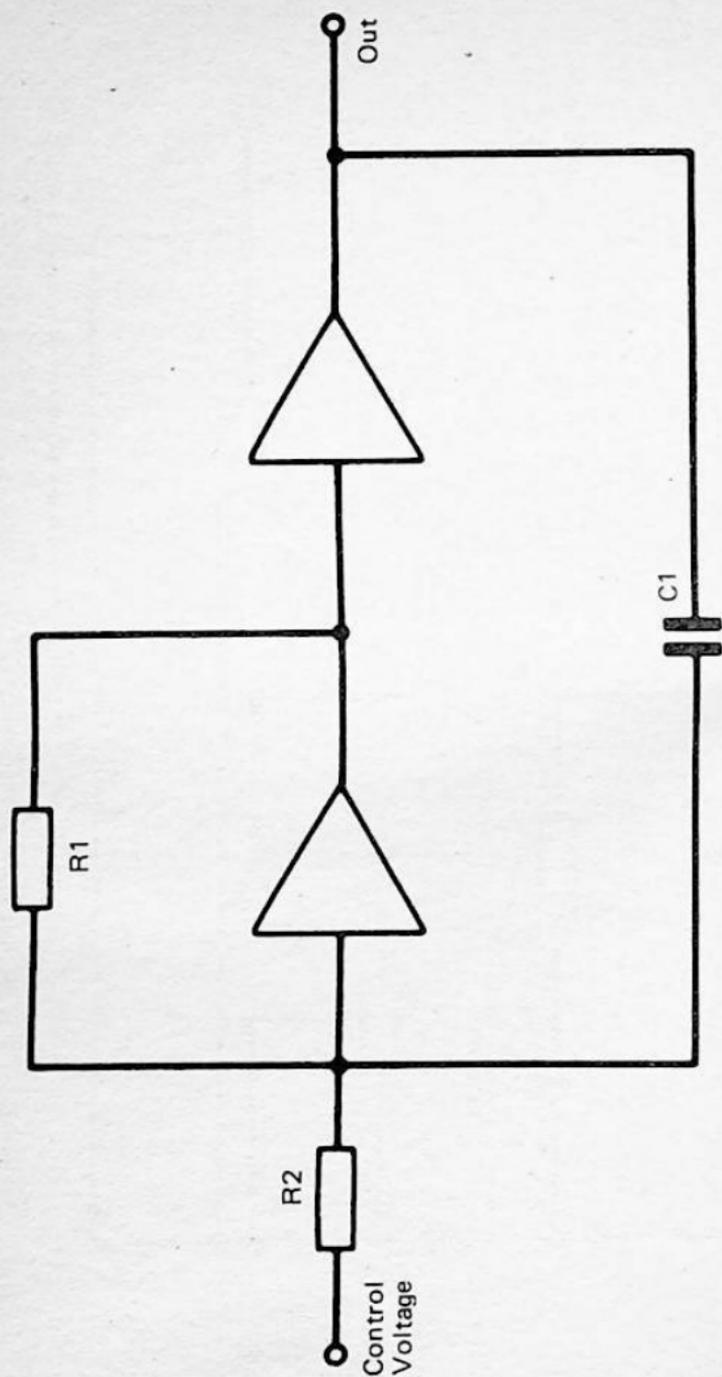


Fig. 13. A simple CMOS V.C.O. circuit

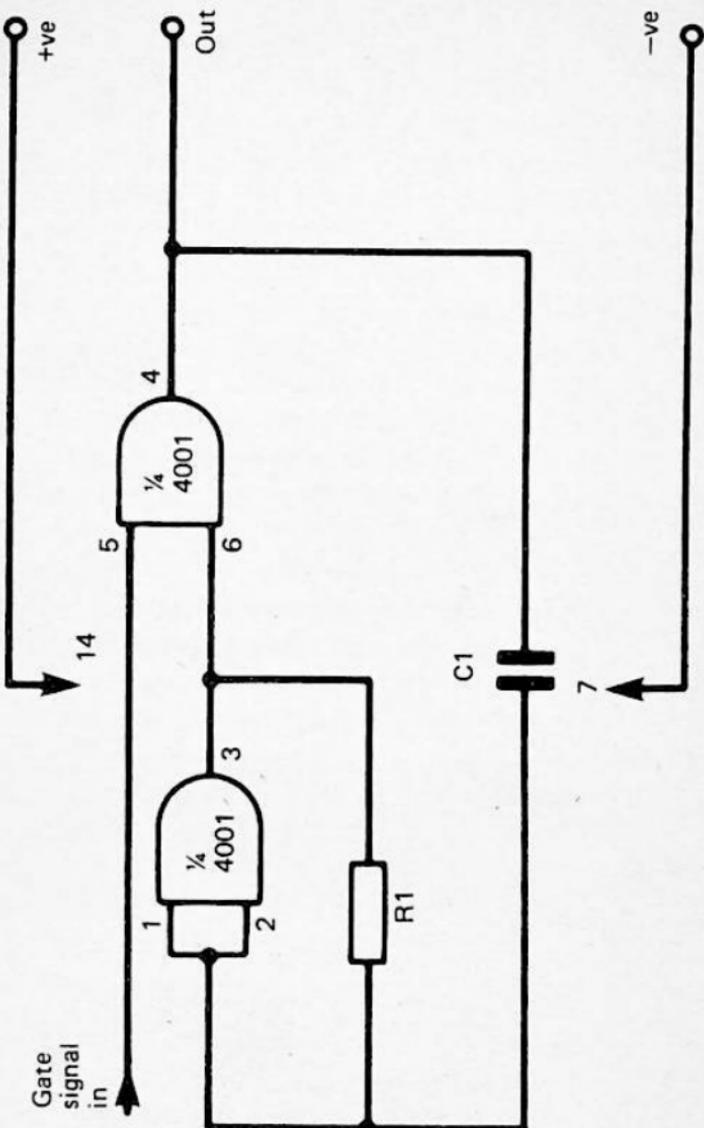


Fig. 14. A gated CMOS astable circuit

gate has one input connected to the output of the previous gate and the other input fed with the gating signal.

With the gate input in the high state the output of the circuit must go low, and the circuit does not oscillate. With the gate input taken to the low state the output stage of the oscillator depends on the input state to the other input of the second gate, which gives a straight forward inverter action. Thus the circuit then oscillates normally.

An alternative gated astable circuit is shown in Figure 15, and this differs from the previous circuit in that the gating signal is applied to one input of the first gate, rather than the second gate. This again gives normal operation with the gate input taken low, and no oscillation when the gating signal is high. However, with the circuit inhibited the output takes up the high state instead of the low one.

If a circuit is required that oscillates when the control signal is high, and is inhibited when it is low, one way of achieving this is to add an inverter ahead of the gate input. One of the unused gates of the 4001 IC could be used as this inverter. An alternative would be to use the 4011 quad 2 input NAND gate instead of the 4001 device, which would give the desired circuit action but would in other respects be the same as using a 4001.

With any CMOS logic device, remember that any unused inputs should be connected to the positive or negative supply rail (whichever is more convenient) and should not simply be left floating.

4047 Astable

The 4047 CMOS IC is specifically intended for use as an astable or monostable multivibrator. We will only consider its use in the astable mode here, but various monostable modes of operation are covered in the next section of this book.

Figure 16 shows the 4047 used as a true gating astable, and in this mode the circuit oscillates when the gate signal is high, and oscillation is blocked when this input signal is low. If a free running astable is required the same basic circuit can be used, but pin 5 of the 4047 should be connected to the positive supply rail.

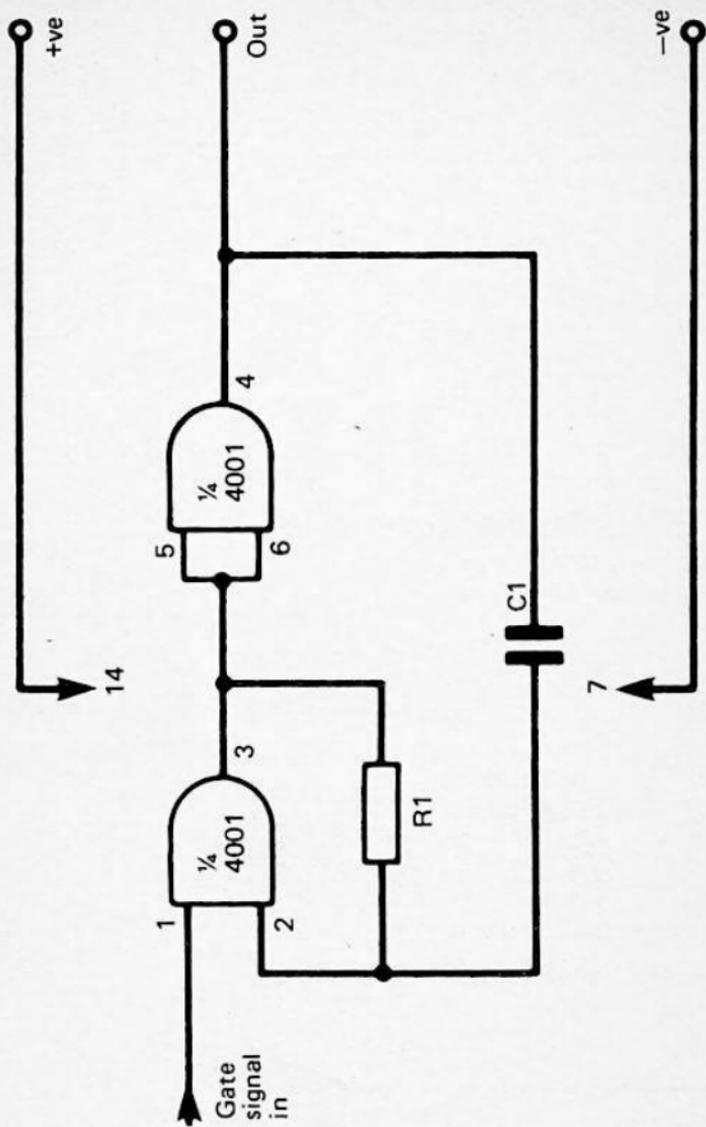


Fig. 15. An alternative gated CMOS astable circuit

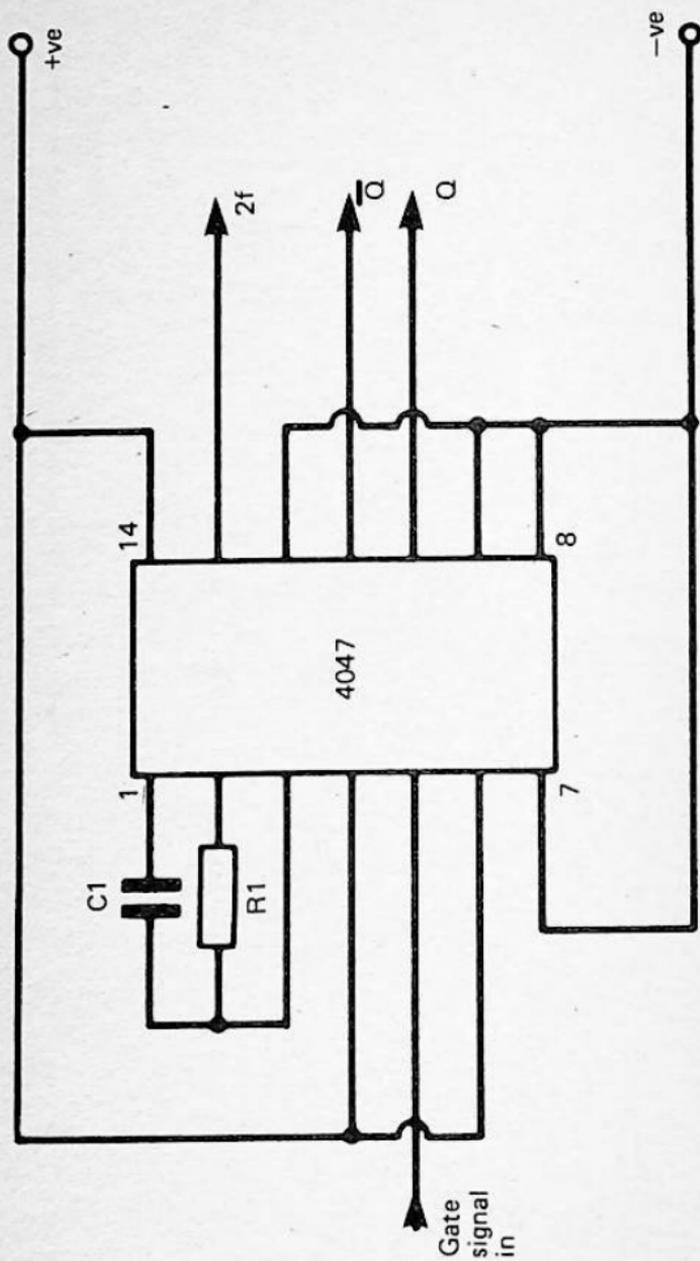


Fig. 16. A 4047 gated or free running astable

The 4047 gives antiphase outputs from the Q and not Q terminals, and this makes it a convenient device to use in applications where a two phase clock signal is required. When gated to the off state the Q output goes low and the not Q output goes high. There is a third output at pin 13 of the 4047, and this gives an output signal which is double the output frequency provided by the main Q and not Q outputs. The waveform at all three outputs is a good quality squarewave.

One output cycle is nominally $4.4RC$ seconds, and the output frequency is nominally equal to $1/4.4RC$ Hz. Timing components R1 and C1 are the only discrete components required, and R1 can have any value from 10k to 1M. C1 can have any value above 100pF, but as it cannot be a polarised type the maximum practical value is about $2\mu 2$.

Complement Gating

The 4047 can also be used as a complement gating astable, where the circuit is blocked when the gate signal is high and enabled when this signal is low. The complement gating astable circuit is shown in Figure 17. In other respects this is the same as the true gating circuit described earlier.

CMOS VCO

A drawback of the 4047 CMOS device is that it cannot be used as a voltage controlled oscillator, but there is another CMOS logic IC, the 4046, which works well in this application. The 4046 is actually a low power, low frequency phase locked loop (PLL), but like any phase locked loop it is based on a voltage controlled oscillator, and by ignoring certain sections of the device it can be used as such.

Figure 18 shows the circuit diagram of a very versatile voltage controlled oscillator using the 4046 device. The 4046 does not have complementary outputs, but a two phase output signal can be obtained, as here, by using one of the devices phase comparators as an inverter to give antiphase outputs. Pin 5 of the device is the inhibit input, and this can be used to gate the oscillator. It is taken high to inhibit the circuit or low to permit

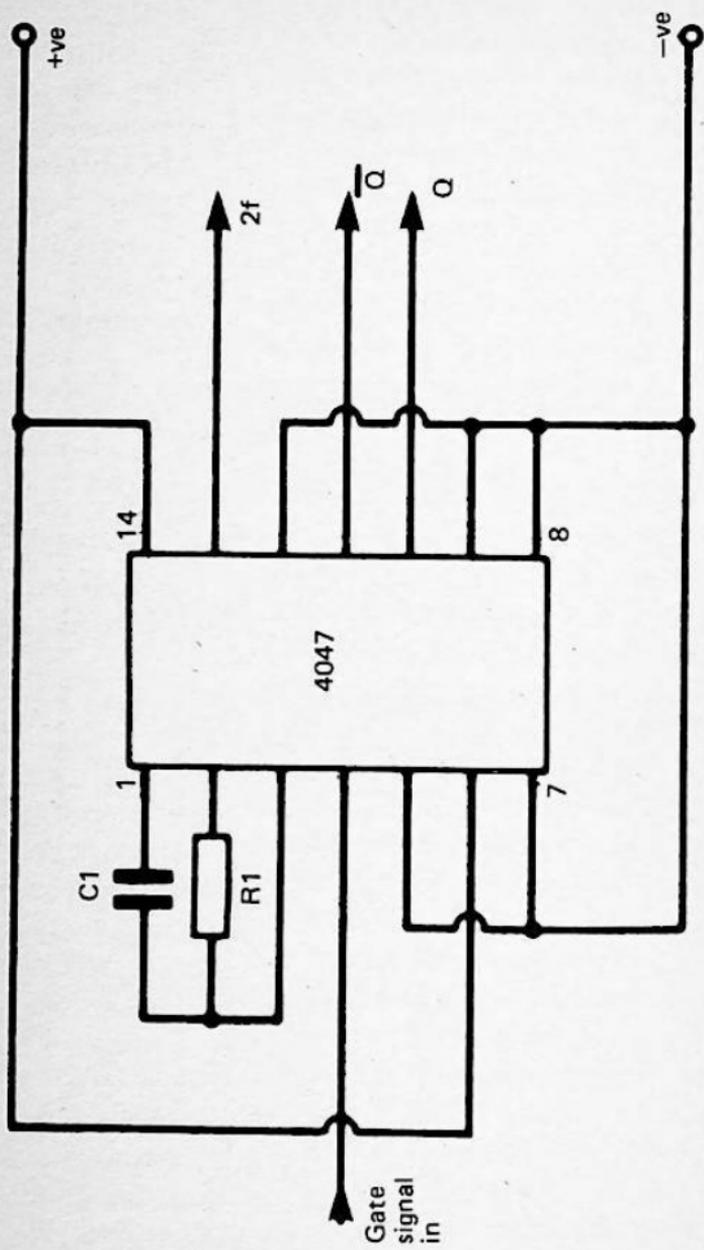


Fig. 17. A complement gating 4047 astable

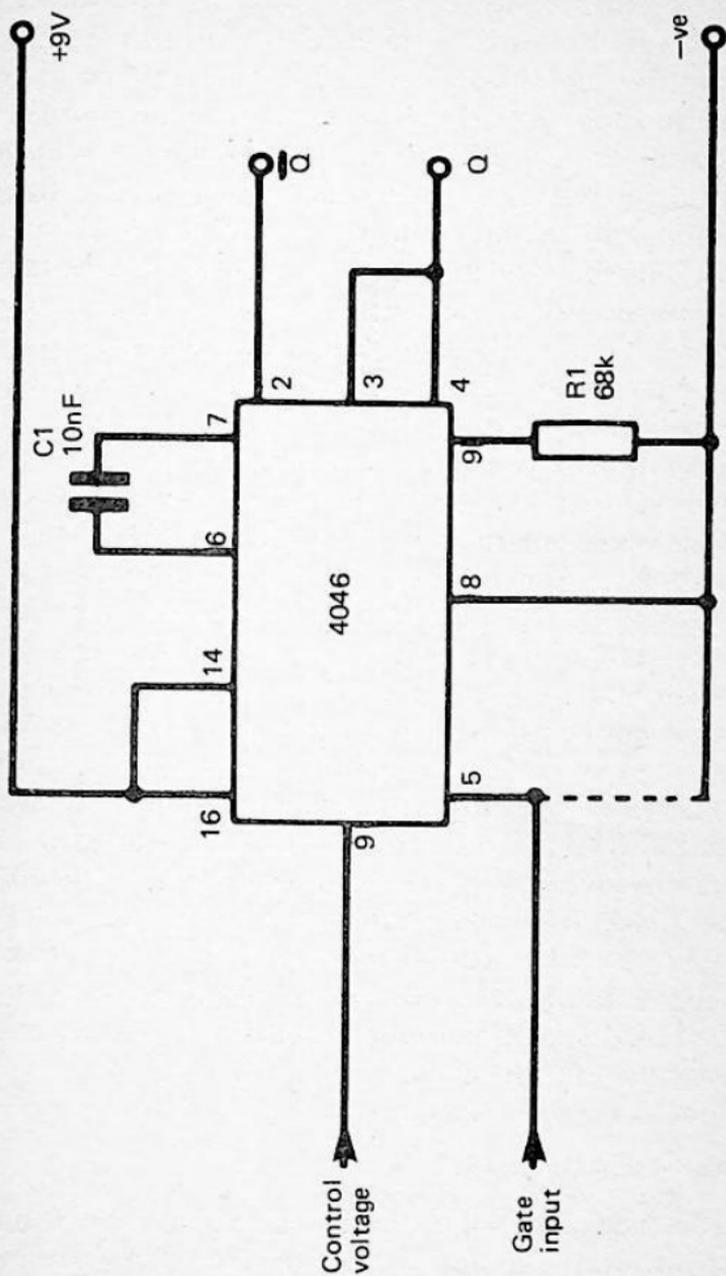


Fig. 18. A simple V.C.O. using the CMOS 4046 device

normal operation. Of course, if this gating facility is not required it is simply necessary to connect pin 5 to the negative supply rail.

C1 and R1 are the timing components, and these set the nominal operating frequency of the circuit at 1kHz with a control voltage of 4.5 volts. Taking the control voltage up to the full positive supply potential of 9 volts gives a rise in frequency to approximately 2kHz, and taking the control voltage down to about 0.6 volts reduces the output frequency to a little under 1Hz. A very wide sweep range is therefore possible using this circuit. If the control voltage is taken below about 0.6 volts the oscillator ceases to function incidentally. The input impedance to the control voltage and gating inputs of the device is extremely high, and no significant current is drawn by either input. The linearity of this circuit is extremely good being typically better than 1%.

The operating frequency of the circuit can be changed by altering the values of C1 and R1, and the circuit will operate from a fraction of one Hertz to more than 500kHz. The output waveform from both the Q and not Q outputs is a good quality squarewave, although at high frequencies there is a slight deterioration of the waveshape. R1 can have any value between 10k and 10M, and C1 can have any value over 50pF (but it must not be a polarised capacitor).

It should be borne in mind that the operating frequency of the circuit will vary slightly with changes in supply voltage even if the control voltage remains constant. In critical applications it is therefore advisable to use a stabilised supply. The 4046, like other CMOS ICs, will operate well over a supply voltage range of 3 to 15 volts. The device has a built-in 5V4 zener diode between pin 15 and the negative supply pin (pin 8), and this can be used as part of a zener stabiliser to produce a regulated supply for the circuit.

Three Waveform Generator

The circuit shown in Figure 19 is based on the 8038CC function generator integrated circuit, and it gives sine, triangular and squarewave outputs. The circuit is very simple, and the only

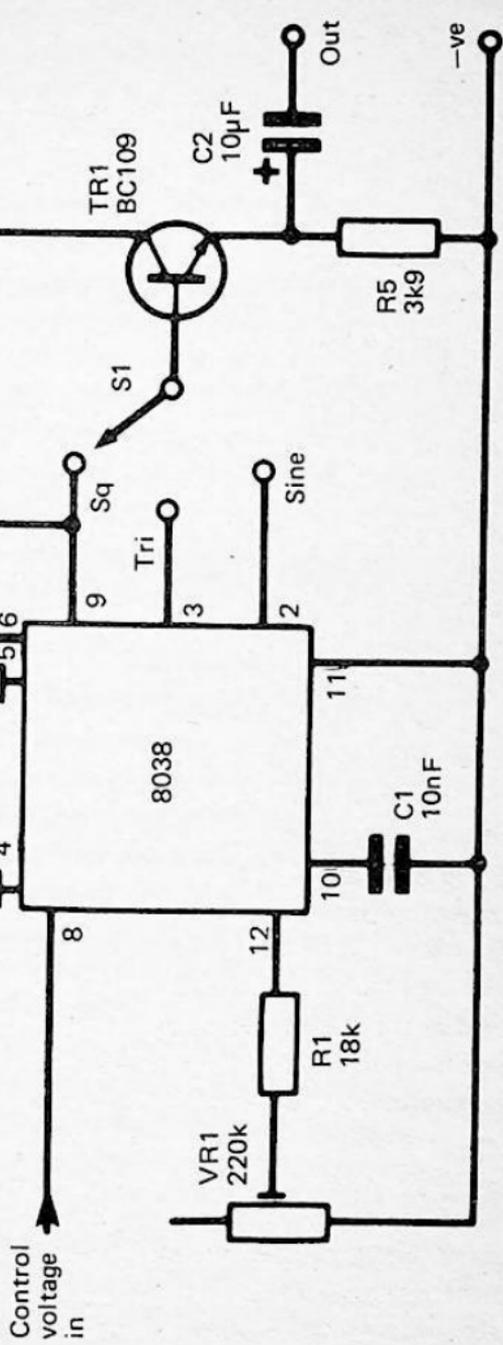


Fig. 19. This V.C.O. offers three output waveforms

real disadvantage it has is the fairly high cost of the 8038CC device when compared with the integrated circuits employed in earlier designs in this book. However, in applications where the three output waveforms available are all required this device probably represents the most cost effective solution to the problem.

Basically the 8038 is a linear triangular waveform generator which also has circuitry to shape this waveform to give square-wave and sinewave outputs as well. One slight drawback of this system is that the sinewave produced is not of high purity, and there is a distortion level of a few percent present on this output signal.

R4 is the discrete load resistor for the squaring circuitry of the 8038, but the other outputs do not require load resistors. The outputs are at a fairly high impedance and a buffer stage must therefore be used at the output. In this case the buffer stage is a simple common emitter stage using Tr1, and S1 is used to select the desired output of the 8038 and connect it to the input of Tr1.

VR2 is used to set the mark-space ratio of the output signals at one-to-one, and VR1 is adjusted to optimise the purity of the sinewave output. These can be adjusted with the aid of an oscilloscope, or it is possible to monitor the output using a crystal earphone and then adjust both components to minimise harmonics on the output. The harmonics are a multiple of the fundamental frequency, and with a fairly low output frequency these should be clearly heard above the fundamental frequency. With some loss of sinewave purity VR1 and R1 can simply be omitted if preferred.

With the specified values the circuit oscillates at approximately 200Hz with the control voltage at the full positive supply potential of 18 volts. Reducing the control voltage results in the output frequency increasing with an output frequency of about 2kHz being reached when the control voltage is reduced by around 1.5 volts. Continuing to decrease the control voltage gives an output frequency of around 10kHz with a control voltage of 12 volts, and a lower control voltage causes oscillation to cease. This circuit therefore gives a reasonably wide control range.

C1 is the only discrete timing component, and the operating

frequency range of the circuit can be altered by using a different value here with changes in value giving an inversely proportional variation in frequency. C1 can have any value of more than about 30pF and a polarised component can be used (the positive leadout connects to pin 10 of the 8038 and the negative lead connects to the negative supply rail). This enables very low frequency operation to be achieved, and the upper frequency limit of the circuit is about 1MHz.

The amplitude of the output signals of the 8038 are not identical incidentally, and the peak to peak output voltage at the squarewave output is virtually equal to the supply potential, while that at the sinewave output is only 0.22 of the supply voltage. The triangular output provides a peak to peak voltage of about 0.3 of the supply voltage.

The circuit will operate with a supply voltage of less than 18 volts, and a supply voltage as low as 9 volts seems to be adequate. However, a lower supply potential does seem to give some reduction in performance. The supply current is about 16mA from an 18 volt supply.

Fixed Frequency Operation

The 8038 can be used in fixed frequency applications, and pin 8 of the device (the sweep input) can then be connected to pin 7. The latter provides an output voltage equal to 0.8 of the supply voltage, and this is used as the control voltage. With C1 at 47nF the circuit has an operating frequency of about 1kHz, and changes in value give an inversely proportional change in frequency. When used in this way the operating frequency is very stable incidentally, and quite large variations in supply voltage do not produce any significant change in output frequency.

NE566N VCO

The NE566N is another useful device for use in voltage controlled oscillator circuits, but it has only squarewave and triangular outputs, and does not have a sinewave output. However, in many applications a squarewave or triangular

output waveform is all that is required.

Figure 20 shows the circuit diagram of a simple VCO using the NE566N, and this circuit has an operating frequency of approximately 1kHz with a 10 volt control voltage. The output frequency reduces to only about 5Hz if the control voltage is taken to around 0.1 volts or so below the positive supply voltage, and oscillation ceases if the control voltage is taken nearer to the supply voltage than this. Taking the control voltage down to about 8 volts boosts the output frequency to around 1.5kHz, but reducing the control potential much further than this results in oscillation ceasing.

Although a 15 volt supply is specified in Figure 20 the circuit will operate with a supply potential anywhere in the range 10 to 24 volts. The supply current is about 7mA.

The output amplitude using a 15 volt supply is about 2.5 volts and 6 volts peak to peak from the triangular and square-wave outputs respectively.

It is, of course, possible to alter the output frequency by changing the values of the timing components (R1 and C1). C1 can have any value that gives an output frequency of no more than about 1MHz but R1 should be between 2k and 20k. C1 can be a polarised type, and in this case the positive leadout wire is connected to pin 7 of the NE566N.

C2 is needed to give good stability and should not be omitted.

Crystal Oscillator

In certain applications such as calibrators and some logic circuits it is necessary to have a highly stable radio frequency oscillator, and these normally have a crystal as the frequency controlling element. Crystals tend to be somewhat more expensive than the alternatives, but give a level of accuracy and stability that is not easily achieved in any other way.

Apart from specialised applications, in certain types of communications equipment for example, an oscillator of this type will normally operate between about 500kHz to 5MHz or so, and Figure 21 shows the circuit of a simple crystal oscillator which will operate over this range.

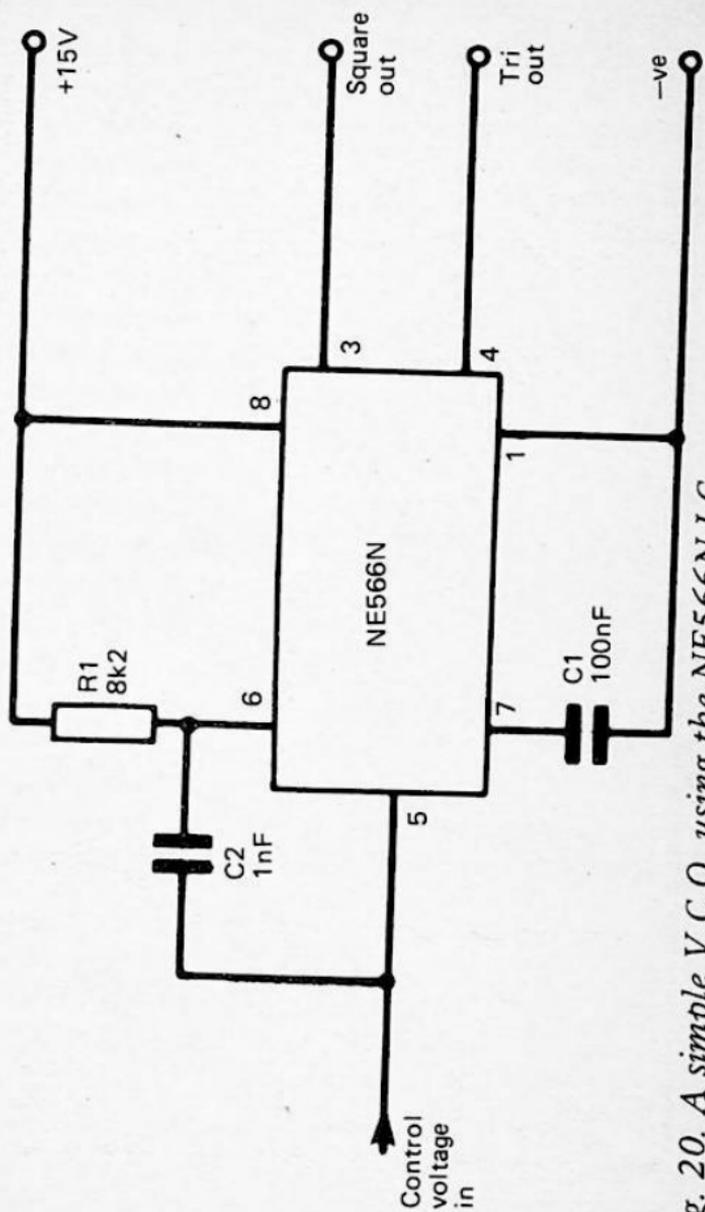


Fig. 20. A simple V.C.O. using the NE566N I.C.

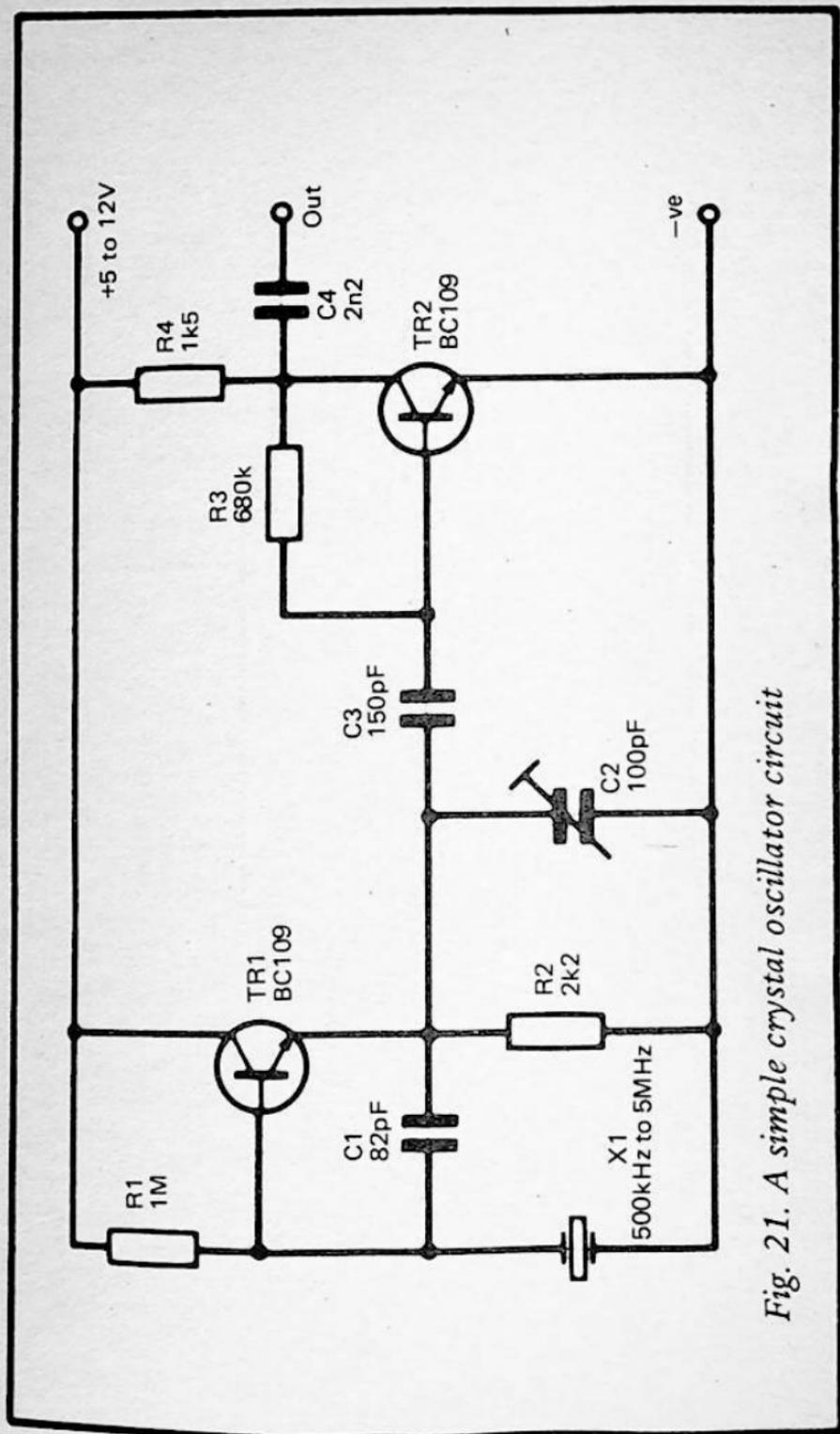


Fig. 21. A simple crystal oscillator circuit

Tr1 is used as a common emitter amplifier, and positive feedback is provided over this stage by crystal X1 and the capacitive tap formed on X1 by C1 and C2. By feeding the output of Tr1 into this tapping the necessary voltage step-up between the emitter and base of Tr1 is achieved, and oscillation is produced at the parallel resonant frequency of the crystal. It is necessary to use a parallel resonant crystal or the circuit will oscillate about two or three kilohertz off-tune, but crystals for use in the 500kHz to 5MHz frequency range are usually meant for parallel resonant operation. C2 can be adjusted to bring the operating frequency of the circuit to within one hertz of the correct frequency if suitable frequency measuring equipment is available. It will not usually be too important if such test equipment is not available since the frequency error is not likely to be more than a few hertz whatever setting is given to C2. This component can be any type having a maximum value of about 100pF, or it can be replaced with a fixed value capacitor of about 56pF in value if the frequency trimming facility is not required.

Tr2 is used as an amplifier which boosts the output signal level to a peak to peak value almost equal to the supply voltage, and it also acts as a buffer amplifier which prevents loading on the output of the circuit from "pulling" the oscillators frequency slightly off-tune. In some applications this amplification and buffering may be superfluous, and Tr2 plus its associated components can then be omitted, with the output being taken from the emitter of Tr1 via C3.

The current consumption of the circuit is about 6mA from a 9 volt supply.

CMOS Crystal Oscillator

Figure 22 shows the circuit diagram of a 500kHz to 5MHz crystal oscillator which uses a couple of CMOS inverters or gates connected to act as inverters. This is in many ways similar to the previous circuit, and the notes for that circuit in general apply to this one as well. The current consumption is only about half that of the previous circuit though.

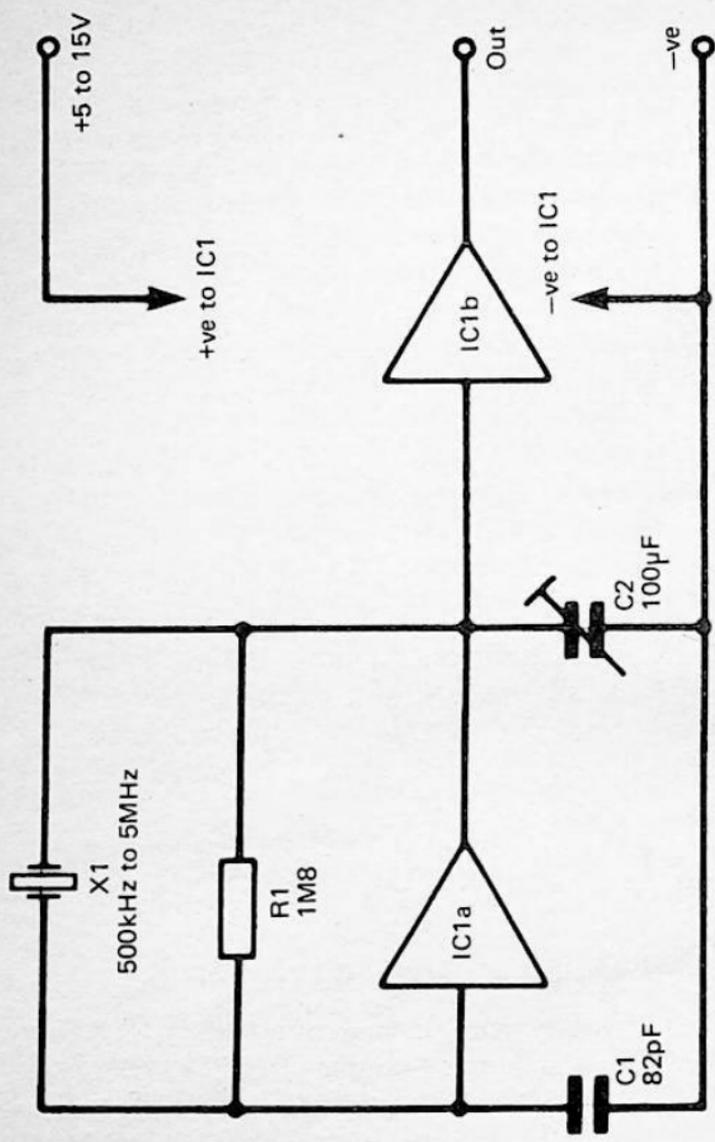


Fig. 22. A CMOS crystal oscillator circuit

L - C Oscillator

While crystal oscillators are ideal for fixed frequency applications, they have the obvious disadvantage of being unsuitable for variable frequency oscillator (VCO) applications. In some cases they can be inconvenient for fixed frequency use due to difficulties in obtaining a suitable crystal. In such cases an L - C oscillator is usually a good solution to the problem, and oscillators of this type can be tuned using either a variable capacitance or a variable inductance (the latter being known as permeability tuning). In virtually all amateur designs the tuning is accomplished using a variable capacitor since permeability tuning can be difficult to arrange mechanically. However, most radio frequency coils have an adjustable core which enables the desired tuning range to be accurately set up, or the desired output frequency to be set if a fixed frequency oscillator is required.

The circuit shown in Figure 23 is for a simple L - C RF oscillator which will operate well from about 100kHz to 50MHz or more. Tr1 is used as a source follower, and the small coupling winding on T1 is used to provide positive feedback and a voltage step-up which sustains oscillation. The main winding on T1 plus C3 are the L - C tuned circuit which determine the operating frequency of the circuit. R1 is the source load for Tr1 at DC, but it is bypassed at AC by C2 so that the coupling winding of T1 is the true source load for Tr1. The output signal is taken from the source of Tr1 by way of DC blocking capacitor C1. C4 is simply a supply decoupling capacitor.

If a tunable oscillator is required, C3 would of course be a variable capacitor, or a combination of fixed and variable capacitors to give the desired tuning range (assuming that permeability tuning is not used). For medium and high frequency oscillators most constructors use ready-made coils these days, and these generally give better and more predictable results than simple home-wound coils. This circuit works well using the Denco transistor useage coils, either the blue aerial type or the yellow RF type. The pin numbers shown in Figure 23 refer to these coils, but note that pins 8 and 9 seem to need the opposite method of connection with the range 5T coils.

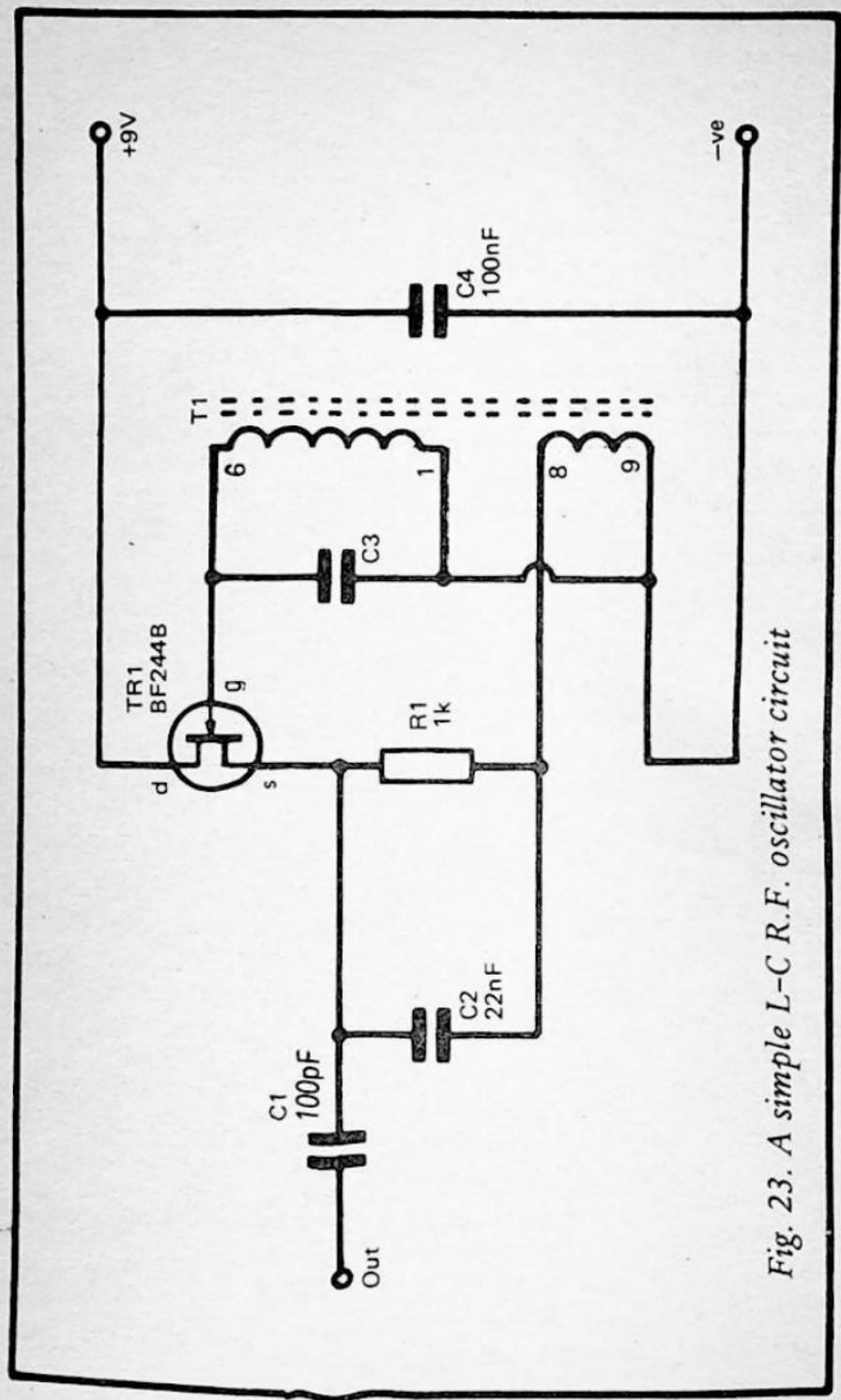


Fig. 23. A simple L-C R.F. oscillator circuit

This circuit is not critical, and there are many other coils, including most single tuned IFTs, which will operate correctly in this circuit. The main point to bear in mind is that the phasing of the small coupling winding must be correct if the circuit is to oscillate, but the correct method of connection can be found by trial and error if necessary.

If the component used in the T1 position has a third winding (as do the Denco coils mentioned above) the output signal can be obtained via this winding instead of by way of C1 if this is more convenient for some reason.

The current consumption of the circuit is only about 1mA or so. Variations in the supply voltage will cause small changes in output frequency, and for optimum stability a stabilised supply should be used.

Single Coil Oscillator

Figure 24 shows the circuit of a useful oscillator which will operate well over a frequency range of about 0.5 to 50MHz using suitable tuning inductors and capacitors, and it has the advantage over the previous circuit of requiring a single tuning coil, rather than a two winding transformer. It does have one slight disadvantage though, and that is a relatively narrow tuning range for a given tuning inductance.

The oscillator is based on Tr1, and Tr2 is merely used as an emitter follower buffer stage at the output. Tr1 is used in the source follower mode, and C3 plus C4 form a capacitive tapping on tuned circuit. The output from the source of Tr1 is coupled into this tapping while the gate of Tr1 is fed with the voltage developed across the tuned circuit, and this gives the voltage step-up and positive feedback needed to produce oscillation at the operating frequency of the tuned circuit.

The input capacitance of Tr1, plus the greater capacitance formed by the series capacitance of C3 and C4, means that the minimum capacitance across tuning coil L1 is going to be something in the region of 40pF. C2 is the main tuning capacitor, and can have a value of anything from zero to around 400pF. The problem with using a higher value here is that it tends to make the capacitive tapping inefficient, and oscillation is likely

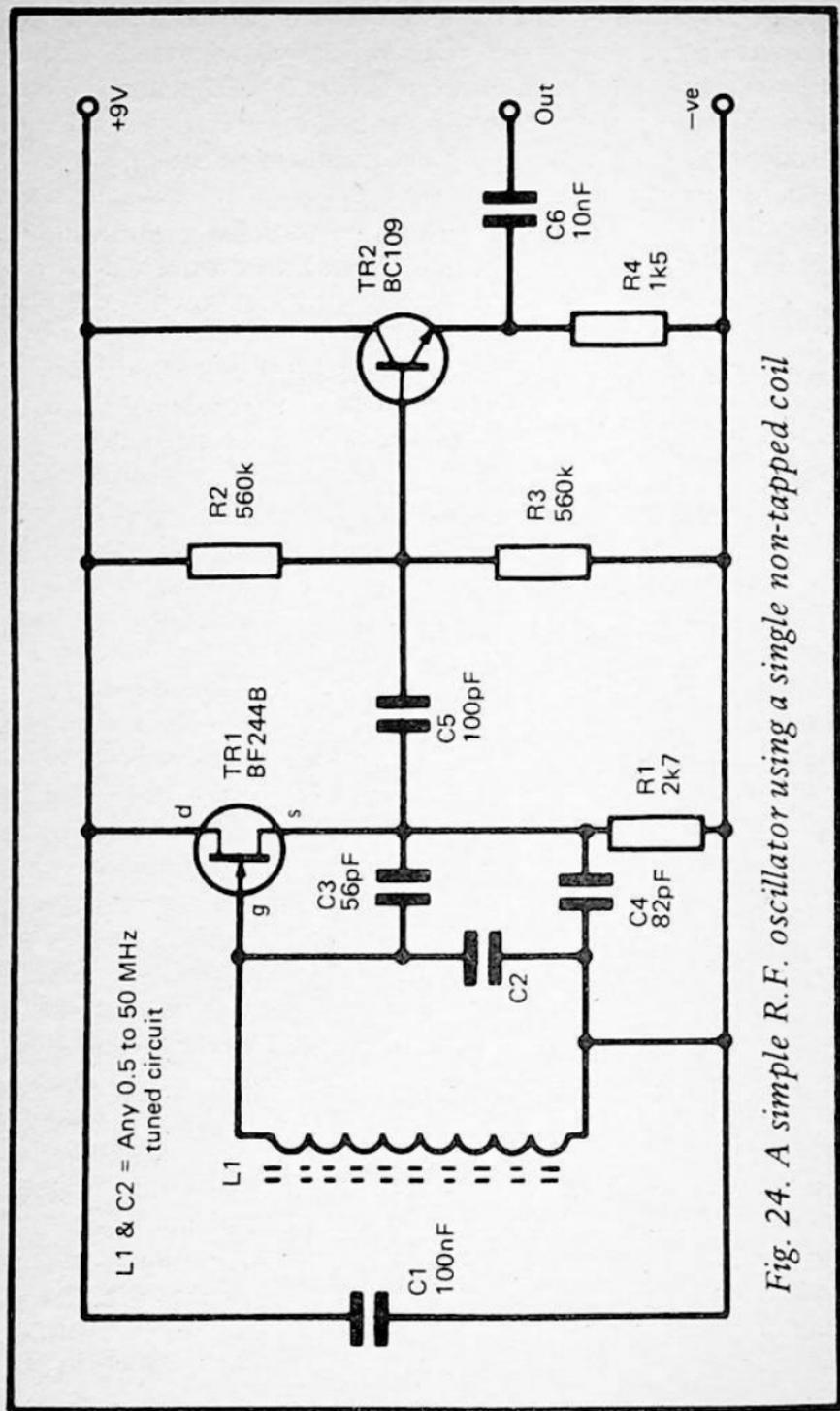


Fig. 24. A simple R.F. oscillator using a single non-tapped coil

to cease. This fairly high minimum capacitance (which will be boosted still further by the minimum capacitance of C2 if this is a variable type) plus the limited although still fairly high maximum capacitance produces the limited tuning span for a given tuning inductance. However, a reasonably wide tuning range is nevertheless possible, and this circuit will normally be adequate in this respect.

Overtone Oscillator

The oscillator circuit shown in Figure 25 has both an L - C tuned circuit (C2 and L1) and a crystal (X1), and it is an overtone oscillator. This is really just another form of crystal oscillator, and although adjusting the tuned circuit values does give some control over the output frequency, the degree of shift that can be achieved in this way is typically only about 2kHz. Thus the circuit is essentially a fixed frequency type having a somewhat wider trimming range than a normal crystal oscillator.

It is called an overtone oscillator because it uses a crystal that is designed to work on a harmonic or overtone, rather than at its fundamental frequency. A harmonic or overtone is simply a multiple of the fundamental frequency, and a third overtone crystal, for example, would operate at a frequency equal to three times its fundamental frequency.

The point of using this method is simply that high frequency crystals tend to be rather fragile, and an overtone crystal therefore tends to be more rugged and practical for high frequency use. It is important to note though, that an overtone crystal will operate at its fundamental frequency if it is used in oscillator circuits such as those shown in Figures 21 and 22 and described earlier. In order to obtain proper overtone operation it is necessary for the circuit to incorporate an L - C tuned circuit resonant at approximately the required overtone frequency so that the circuit works efficiently at this overtone frequency, but is inefficient at the fundamental frequency. This suppresses oscillation at the fundamental frequency and encourages oscillation at the overtone. Even so, the circuit is unlikely to operate properly using an ordinary crystal on an

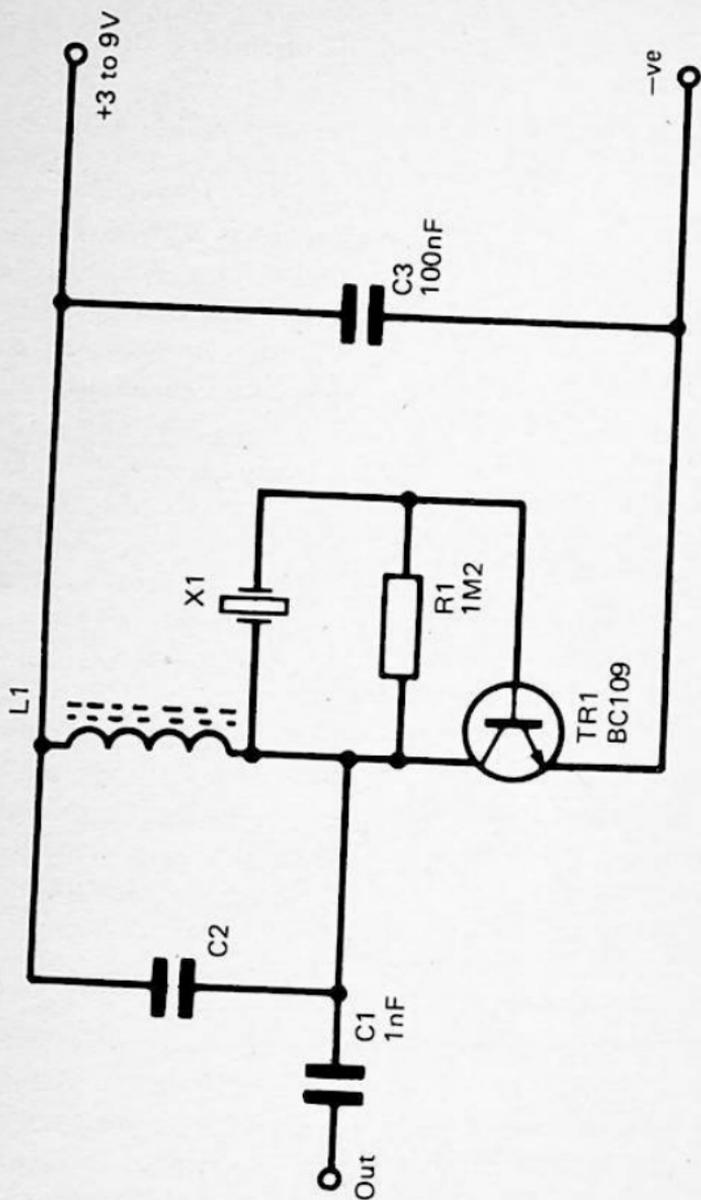


Fig. 25. A crystal overtone oscillator

overtone, and it is necessary to use a proper overtone crystal on the appropriate overtone. Retailers or manufacturers data normally states that a crystal is an overtone type where this is applicable.

Sawtooth Generator

A sawtooth waveform is useful for a number of applications, such as in simple wobblers, sweep generators, and modulated tone generators. To the best of the authors knowledge there is no integrated circuit designed specifically to produce this type of waveform, but it is still quite an easy waveform to produce.

Figure 26 shows the circuit diagram of a simple non-linear sawtooth waveform generator which is basically just a standard 555 astable. However, the resistance between pins 6 and 7 of the device has been made zero so that timing capacitor C1 is discharged almost instantly when its charge potential reaches the $2/3 V+$ threshold level. It is the waveform generated across C1 that is the required sawtooth waveform, and the brief negative pulses at pin 3 of IC1 are unused. Tr1 and R1 form a simple emitter follower buffer stage which gives a reasonably low output impedance from the high impedance signal across C1.

The specified values give an operating frequency of about 1kHz (960Hz in theory), but the values of R1 and C1 can of course be changed to give a different output frequency. It is advisable not to make R1 much higher than the specified value of 150k though, or loading on the timing network could prevent the circuit from functioning properly.

Linear Sawtooth

The waveform produced by the circuit of Figure 26 is a non-linear sawtooth because capacitor C1 charges exponentially via timing resistor R1. In other words, as the charge voltage on C1 increases, the voltage across R1 decreases, and the charge current therefore decreases as well, so that the voltage across C1 rises at a slower and slower rate. A linear sawtooth waveform can be produced by charging C1 from a constant current source so

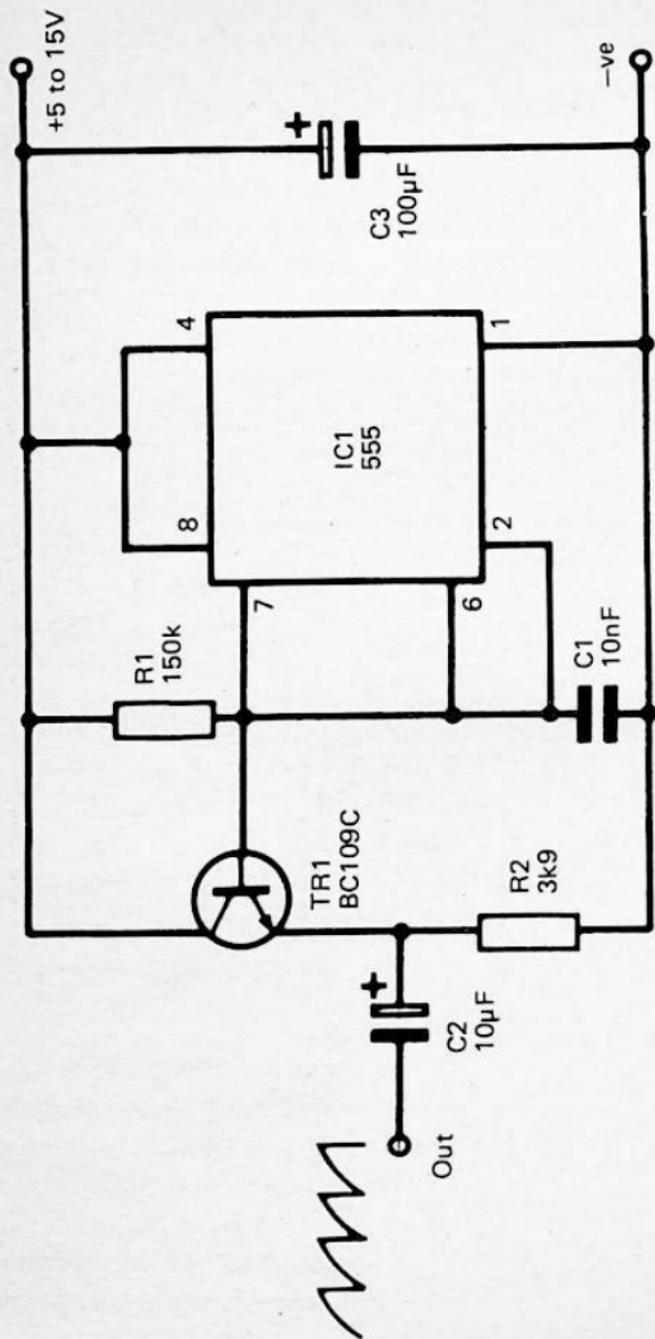


Fig. 26. A simple sawtooth generator circuit

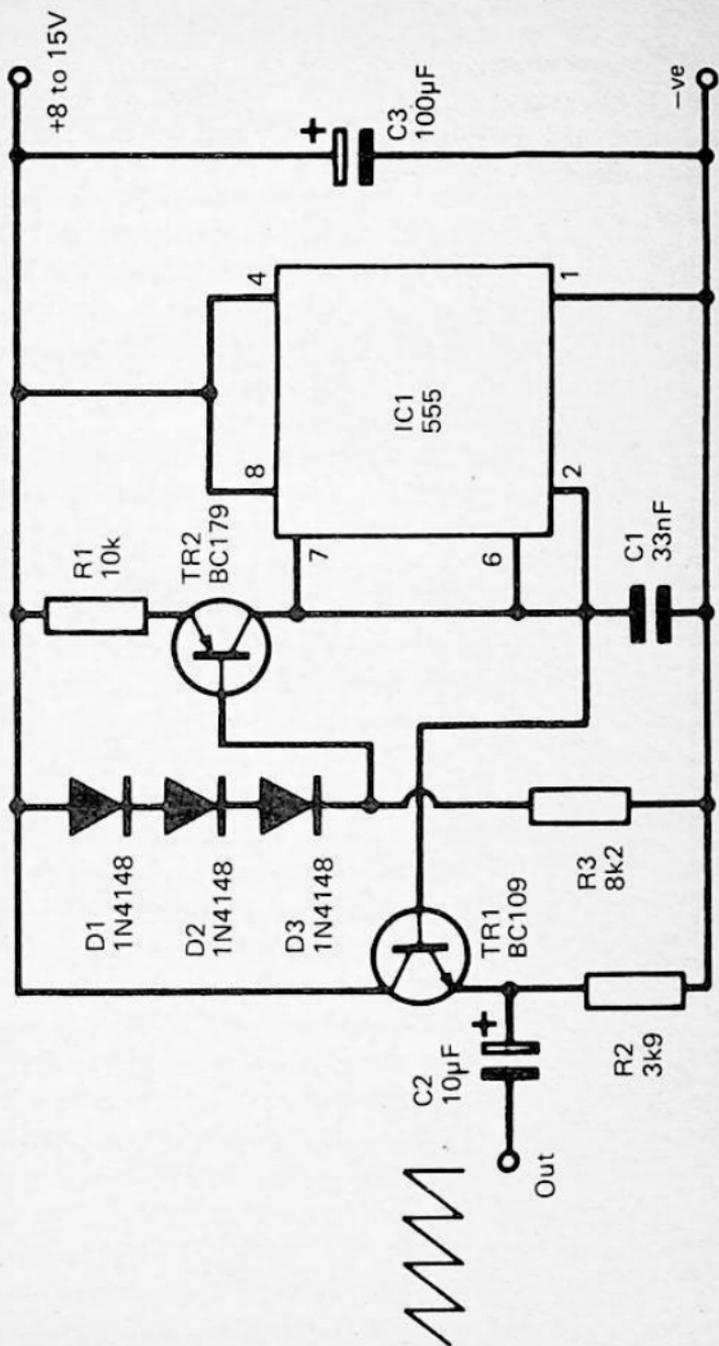


Fig. 27. A simple linear sawtooth generator

that the voltage across C1 rises at a constant rate. Figure 27 shows a simple circuit of this type, and it is the same as the previous circuit apart from the replacement of the timing resistor with a conventional constant current generator based on Tr2.

The circuit operates at about 1kHz using the specified values for R1 and C1, but the output frequency can be altered by changing the value of either of these components. R1 should be kept between about 1k and 47k in value.

Chapter 2

MONOSTABLE MULTIVIBRATORS

A monostable multivibrator produces a single output pulse of a certain duration when it is triggered by an input signal of some kind. Circuits of this type are used in timers, switch debouncers, and many other applications. There are actually several types of monostable multivibrator, and in this chapter we will consider a range of circuits that will cover virtually any requirement, and which will give output times from less than a microsecond to several days.

Basic 555 Monostable

The 555 timer IC can be used as a monostable multivibrator, and is often the most cost effective choice when a circuit of this type is required. Figure 28 shows the basic 555 monostable configuration.

Normally the input of the circuit is held at more than one third of the supply voltage, and under stand-by conditions the internal discharge transistor of the 555 holds C1 in an uncharged state and the output at pin 3 is low. If the trigger input is taken below about one third of the supply potential the circuit is triggered, pin 3 of the 555 goes high, and the internal discharge transistor switches off so that C1 can charge by way of R1. This continues until the charge potential on C1 reaches two thirds of the supply voltage, and the circuit then reverts to its original state. At least, it returns to its original state provided the trigger input has been taken back above one third of the supply voltage. If it has not, the output pulse does not cease until it has been taken back above the one third of the supply voltage threshold level. This is an important point which should be kept in mind.

The output pulse has a duration of $1.1CR$ seconds, and the minimum pulse length is about one microsecond, and the maximum pulse duration is limited only by the maximum practical timing capacitance (as for the astable mode). An

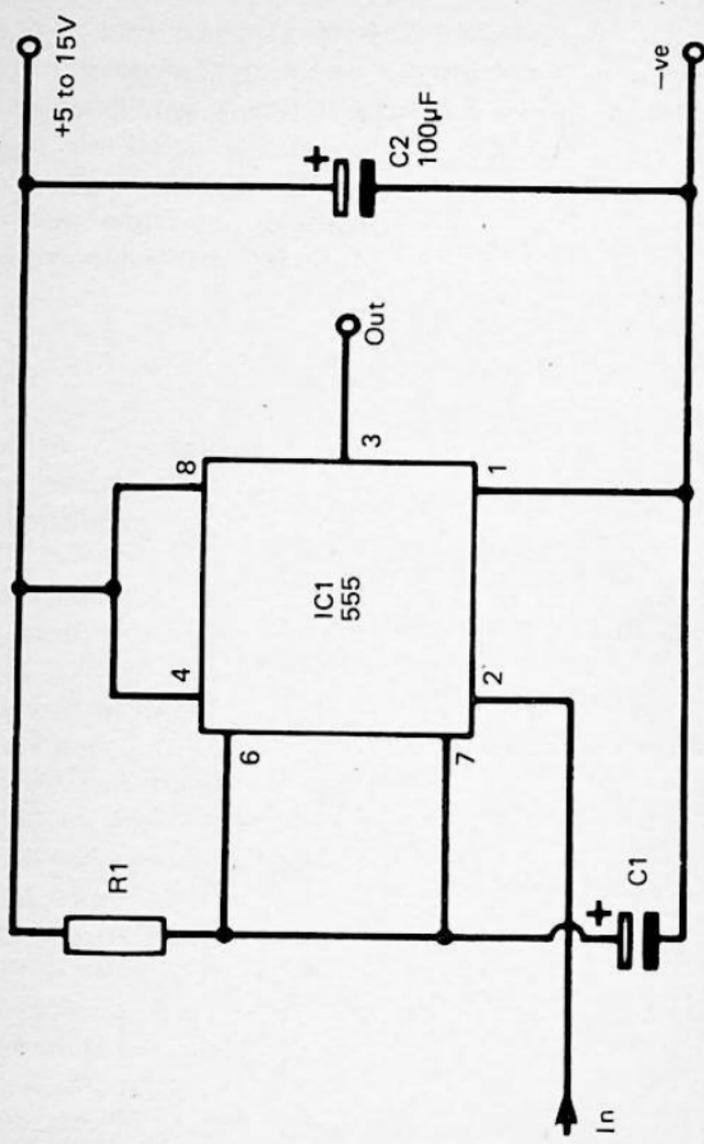


Fig. 28. The basic 555 monostable configuration

output pulse of several minutes can be easily achieved though.

Manual Triggering

Figure 29 shows a very simple method of manually triggering a 555 monostable, but this method is only suitable in cases where an output pulse duration of a few seconds or more is produced by the circuit. PB1 should then be released before the output pulse ceases, and correct operation of the circuit will be obtained. This will not be the case if an output pulse of only a fraction of a second is desired, and it is likely that the pulse duration will be equal to the time that PB1 is depressed rather than the time set by R1 and C1 in such cases.

An improved method of manual triggering is shown in Figure 30. Here R3 normally holds the trigger input at virtually the full positive supply voltage, but a brief negative pulse will be produced at the input when BP1 is operated due to C2 charging via R3. This gives the brief trigger pulse, and no matter how long PB1 is depressed, it is the time constant of R3 and C2 that sets the length of the trigger pulse. R2 rapidly discharges C2 when PB1 is released so that the circuit is almost instantly ready for another operation of PB1.

One slight problem that might arise with this circuit is multiple triggering of the circuit, with IC1 being triggered when PB1 is released. This is due to contact bounce in PB1, and can be combatted by making R2 higher in value so that C2 takes longer to discharge. This will limit the frequency at which the circuit can be triggered, but this is unlikely to be of any practical consequence. The trigger pulse applied to IC1 is only a few microseconds in duration, and the circuit will obviously work properly with output pulse lengths of around $10\mu\text{s}$ or more. For shorter pulse lengths of only about $1\mu\text{s}$ or so it would be necessary to decrease C2 in value, although triggering might become unreliable if this component is made much lower in value.

Reset Facility

In the monostable mode pin 4 of the 555 device is the reset

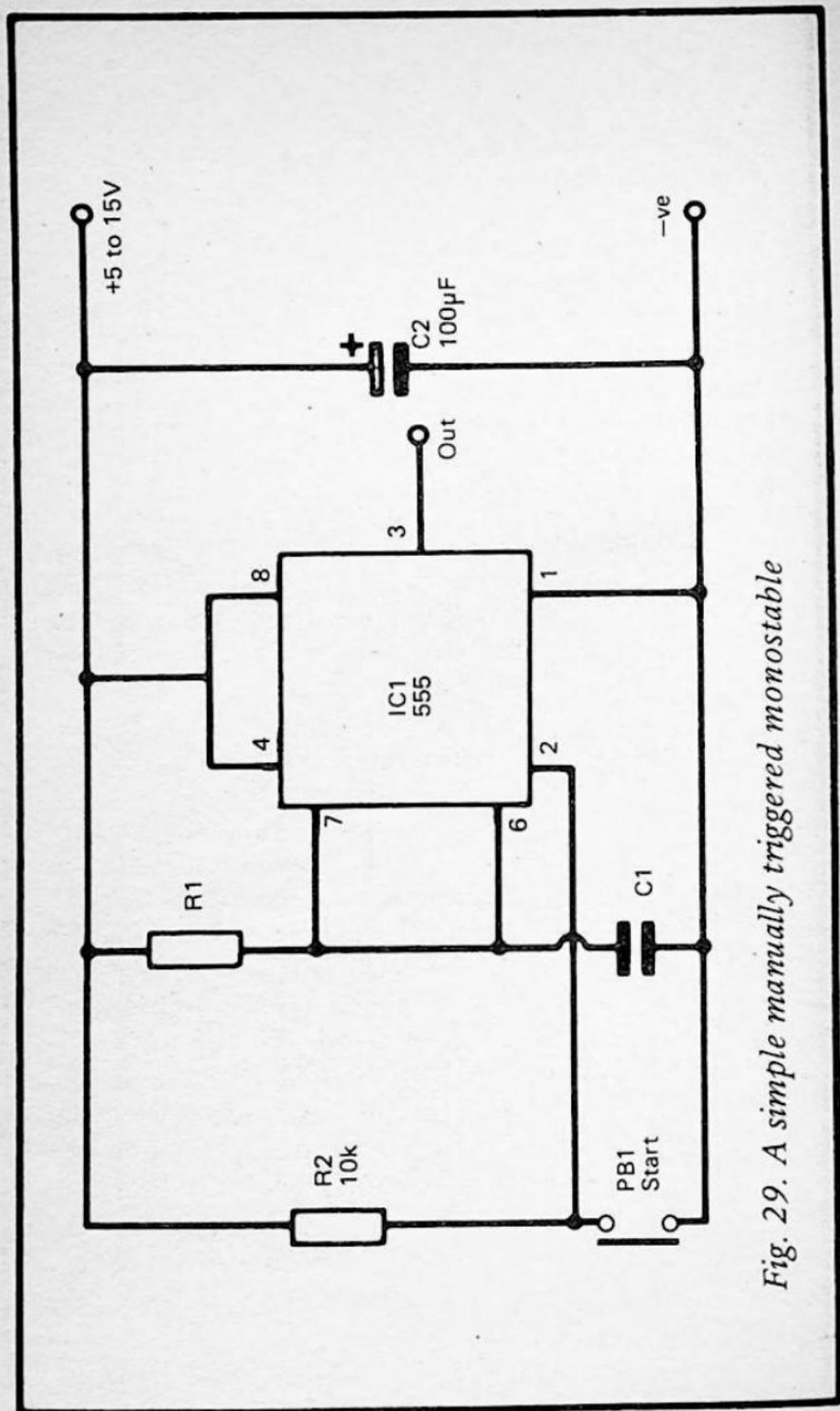


Fig. 29. A simple manually triggered monostable

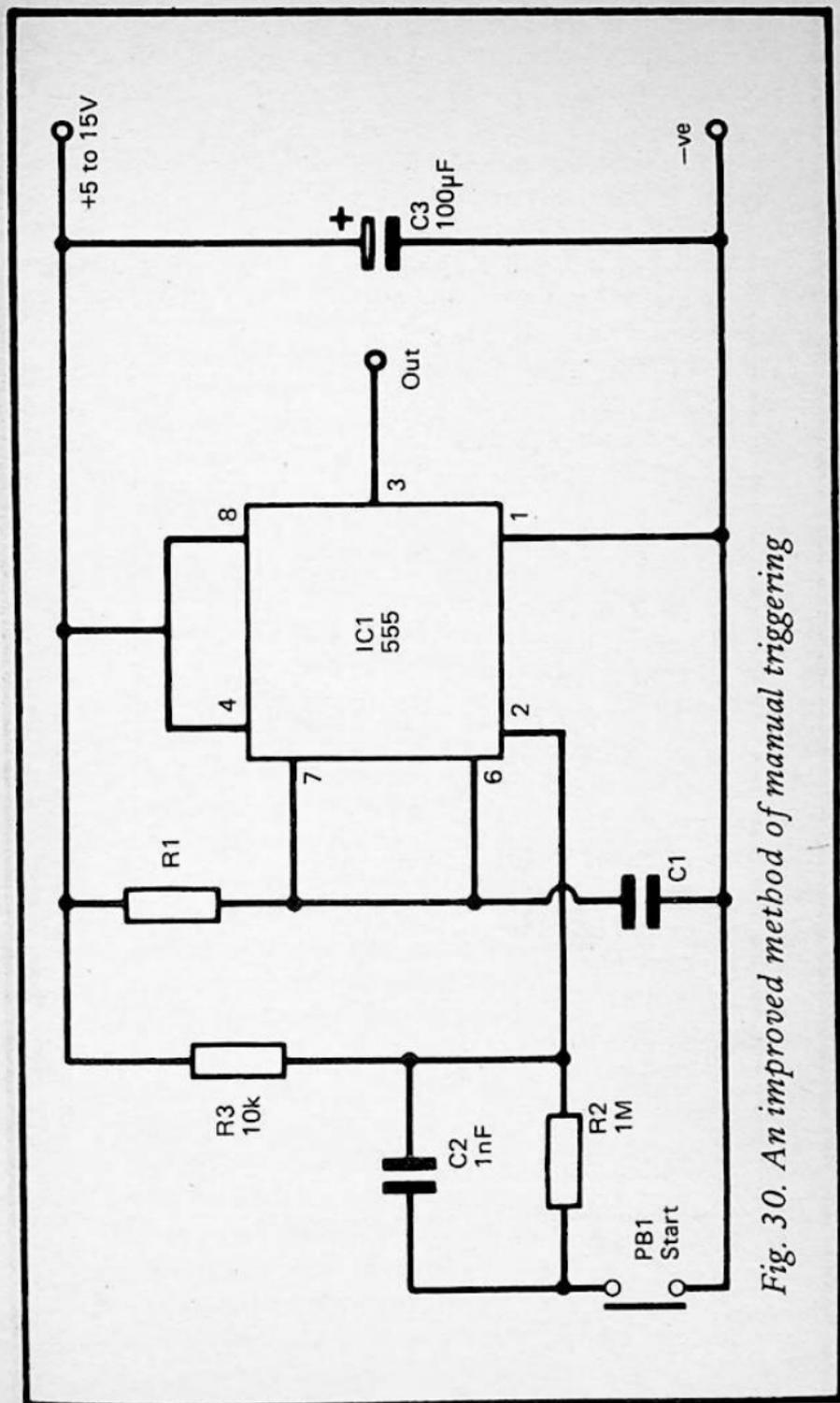


Fig. 30. An improved method of manual triggering

terminal, and this should normally be in the high state. Taking this pin low during the output pulse causes the pulse to be immediately brought to an end. Figure 31 shows how to add both manual and electronic reset facilities to a 555 monostable.

Normally R2 holds pin 4 of IC1 at almost the full positive supply rail voltage, but operating PB1 obviously takes pin 4 low and resets IC1. Similarly, taking the gate terminal of Tr1 to a potential of more than about one or two volts results in Tr1 being biased hard into conduction so that it has a drain to source resistance of only a few ohms, and IC1 is reset.

The reset pulse need only be a few microseconds or more in duration, and must take pin 4 below approximately 0.5 volts. If the electronic reset facility is not required simply omit Tr1, and leave out PB1 if the manual reset facility is not needed.

Pulse Shortener

If a 555 monostable is to be used as a pulse shortener, where the input pulse will still be present when the output pulse should cease, it is obviously not possible to take the input signal direct to the trigger input. Instead it is necessary to use capacitive coupling as shown in the circuit of Figure 32. R2 and R3 bias the trigger input of IC1 to about half the supply voltage so that under quiescent conditions the circuit is not triggered. If an input signal is coupled to pin 2 of IC1 via C2, and this signal is a pulse of fairly long duration, the signal at pin 2 will in fact be a brief negative pulse followed by a brief positive pulse. The first pulse is on the negative going edge of the trigger pulse and the second pulse is caused by the positive going edge of the trigger pulse. This distortion of the trigger pulse is caused by the low value of C2 in relation to that of R2 and R3, and the consequent poor low frequency coupling that is provided.

The positive going pulse is irrelevant, and it is on the initial negative going pulse that the circuit is triggered. As this pulse only lasts a few microseconds no matter how long the input pulse, the circuit works perfectly well as a pulse shortener. However, note that the input signal must have an amplitude of a few volts peak to peak in order to trigger the

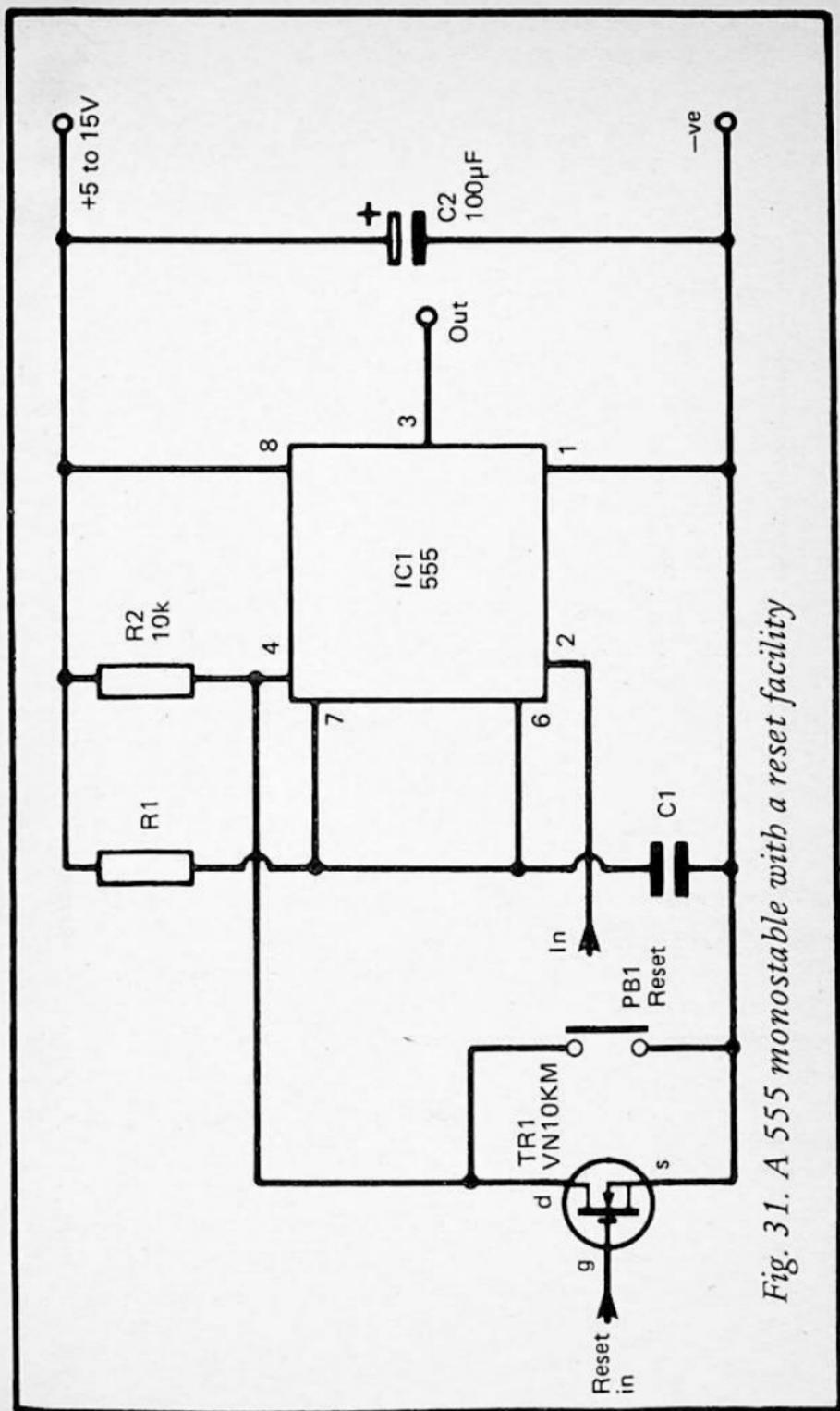


Fig. 31. A 555 monostable with a reset facility

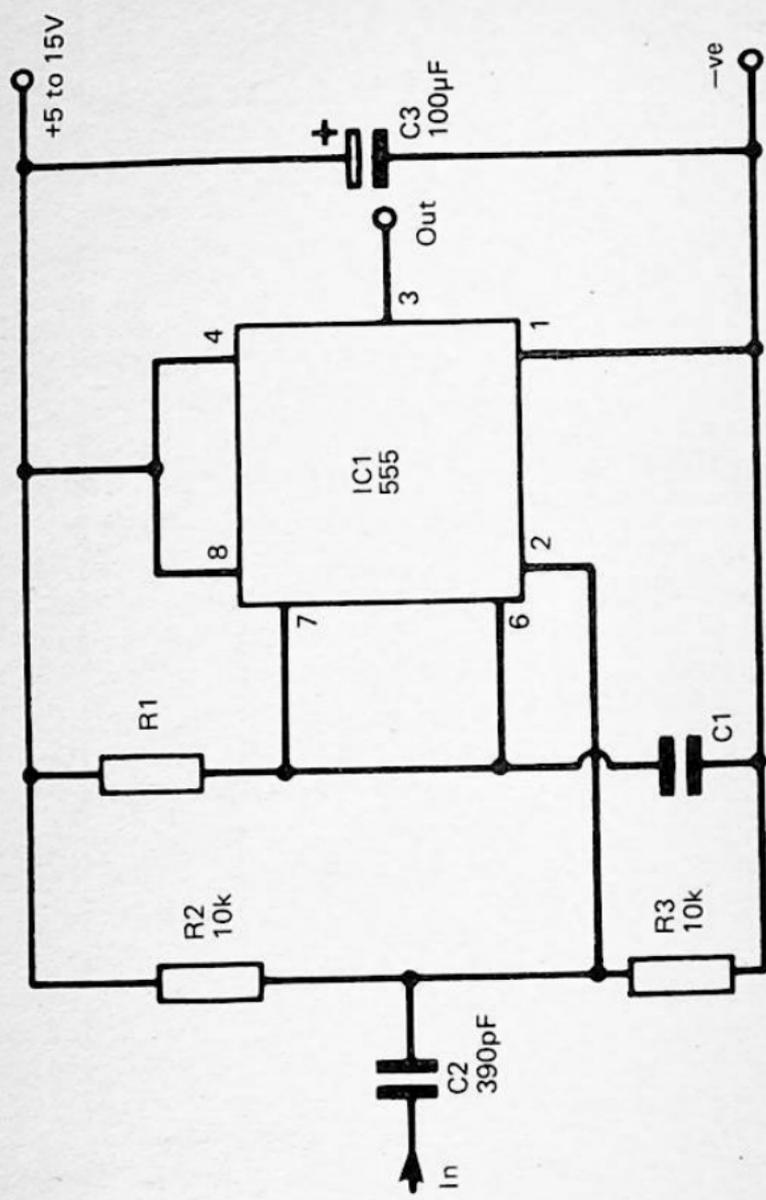


Fig. 32. A 555 pulse shortener

circuit, and that it must have a fast fall-time. The circuit will trigger reliably from a low frequency squarewave, but not from a sinewave signal.

4047 Monostables

The 4047 CMOS IC, as mentioned in the previous chapter, can be used as a monostable multivibrator as well as an astable type. There are actually three monostable modes in which it can be used: positive triggered, negative triggered, and retriggerable. These three modes of operation use the circuits shown in Figures 33 to 35 respectively.

The positive triggered type is triggered as the input is taken from the low state to the high one, whereas the negative type is triggered as the input goes from the high state to the low one. With both these circuits the output pulse ceases after the appropriate period of time even if the input is still in the appropriate trigger state. It is in this respect that the retriggerable type differs from the other two, and, like a 555 monostable, the output pulse of the retriggerable type does not cease while the trigger pulse is present. The retriggerable circuit is a positive edge triggered type incidentally.

With all three circuits the output pulse duration is nominally $2.48 RC$ seconds, and $R1$ should be between $10k$ and $1M$. $C1$ can have any value over $1nF$, but it must not be a polarised type. In all three modes the Q output is normally low and goes high to produce a positive output pulse, while the not Q output is normally high and gives a negative output pulse.

CMOS Monostable

An alternative to using a special CMOS monostable device is to use CMOS gates connected to give a monostable action. This has the advantage of low cost, but admittedly there is the disadvantage that the output pulse length cannot be predicted with quite the level of accuracy that is possible using a special device such as the 4047. However, in many applications this is not a significant drawback, and a simple CMOS monostable is a very useful circuit.

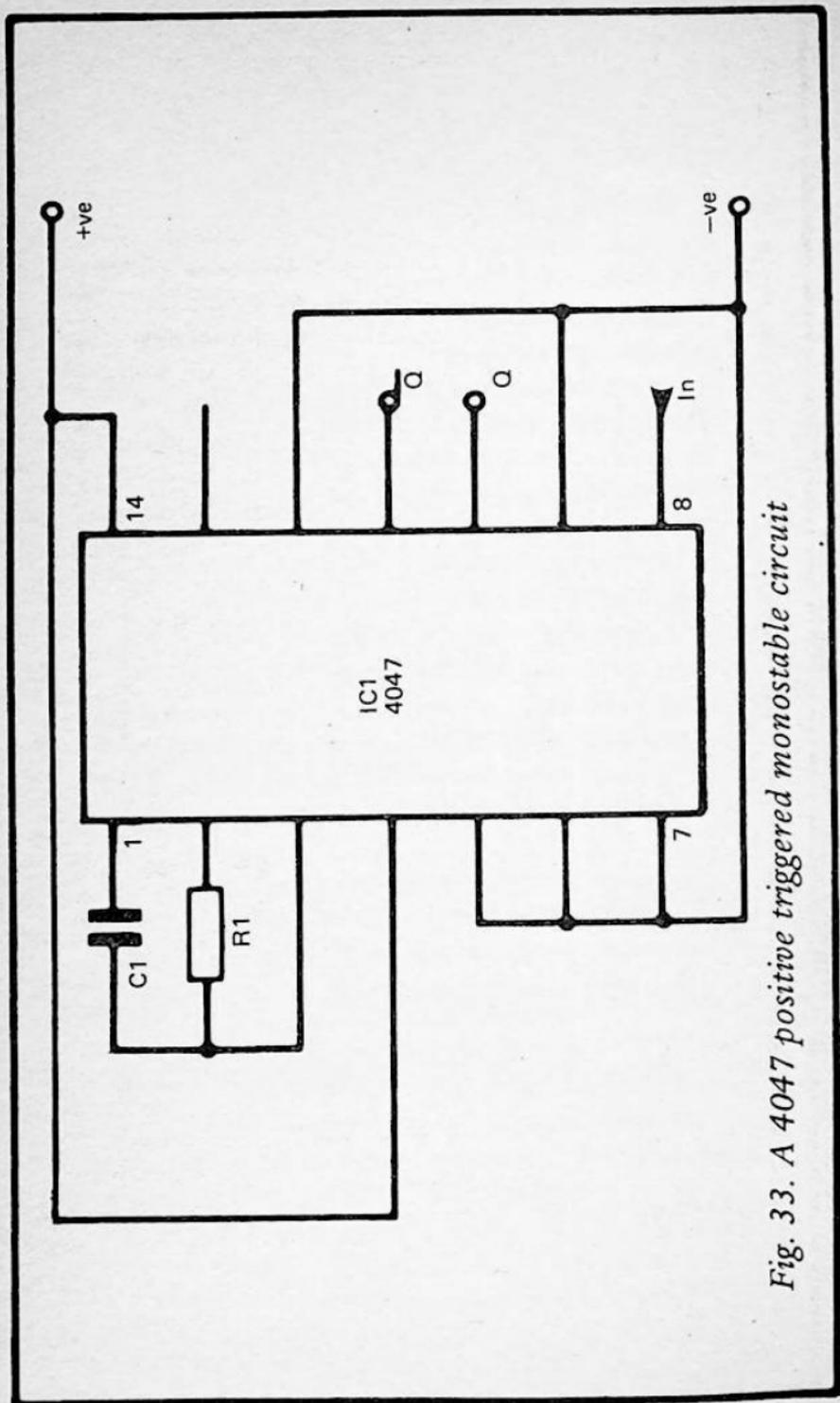


Fig. 33. A 4047 positive triggered monostable circuit

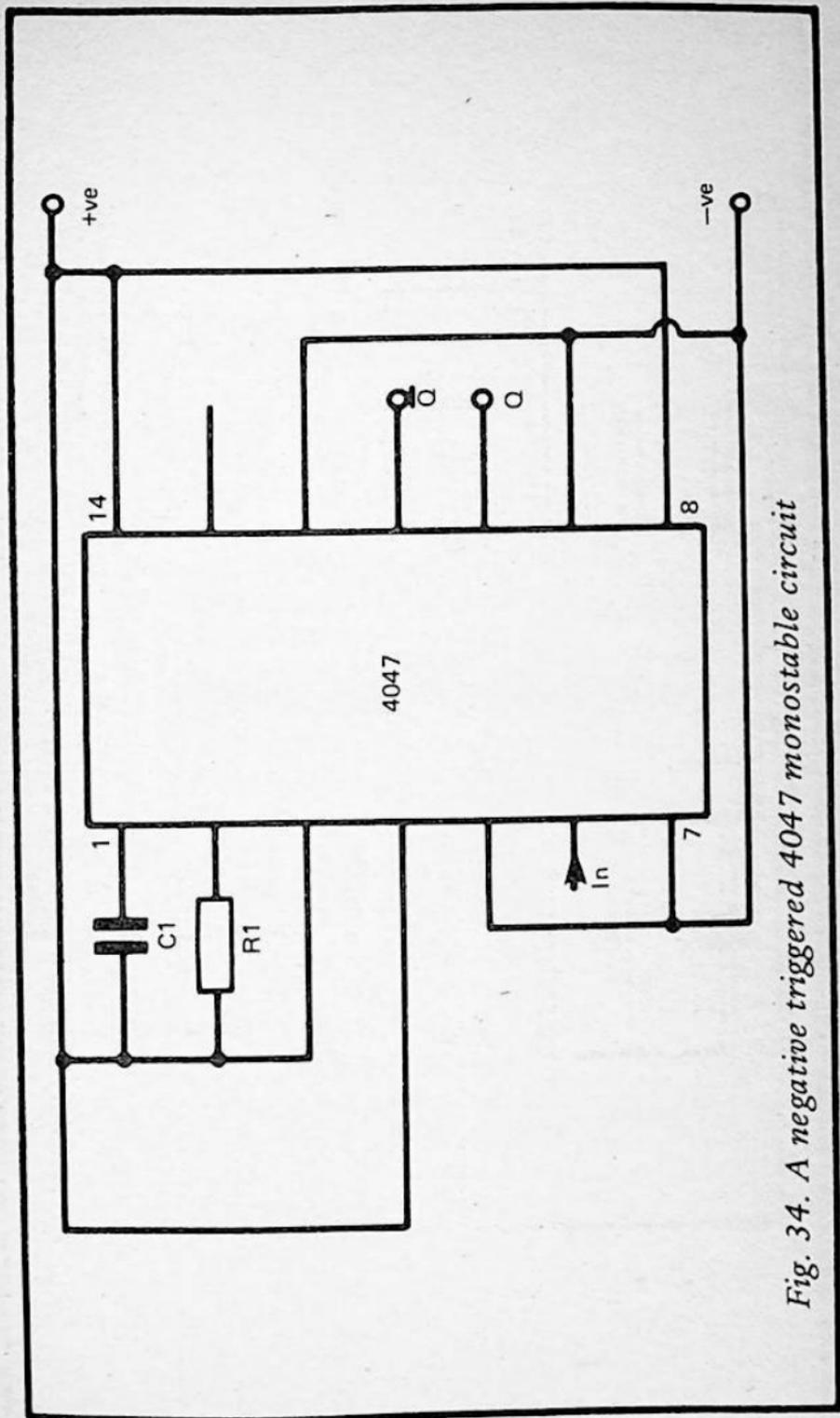


Fig. 34. A negative triggered 4047 monostable circuit

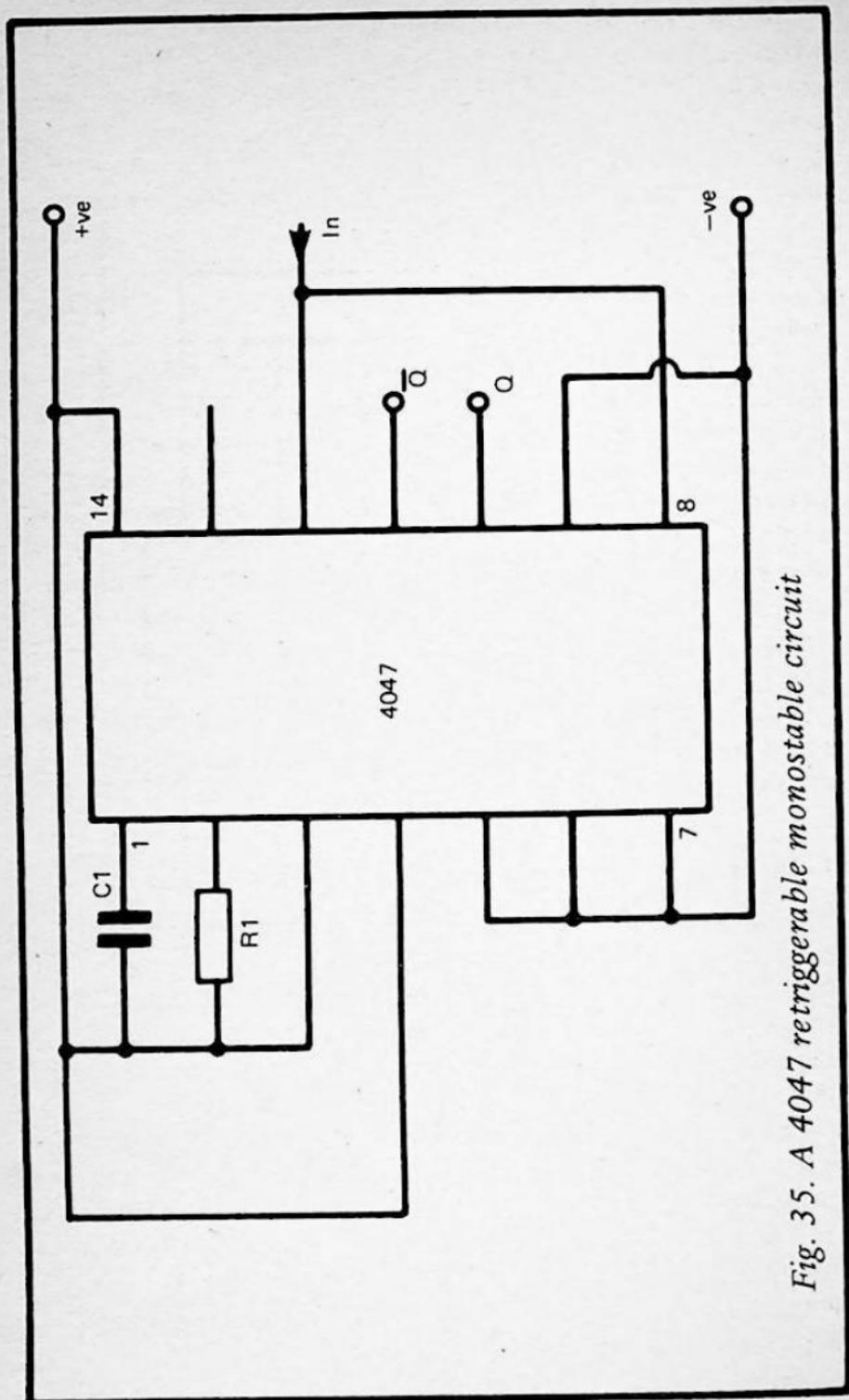


Fig. 35. A 4047 retriggerable monostable circuit

Figure 36 shows the circuit diagram of a CMOS monostable which is based on a couple of the four 2 input NOR gates contained in the 4001 CMOS device. C1 and R1 are the timing components, and C1 can be a polarised type. The correct polarity is shown in Figure 36. R1 can have any value from about 10k to many Megohms, and C1 can have any value of more than about 30pF. The output pulse duration is approximately equal to $0.65 CR$, but it must be stressed that this is only very approximate, and will vary significantly from one 4001 IC to another.

This monostable is a positive triggered type, and the input is therefore normally in the low state and is taken high to trigger the circuit. The output is normally low and gives a positive output pulse. By using the otherwise unused gates of IC1 as inverters it is possible to obtain other modes of operation, and adding an inverter at the input makes the circuit a negative edge triggered type. Adding an inverter at the output gives a negative output pulse in addition to the positive one, giving Q and not Q outputs. These modifications are shown in the circuit diagram of Figure 37.

Note that these monostable circuits are of the type where the input pulse can be longer than the output pulse, and they are not retriggerable types.

Long Timer

The monostables described so far are primarily intended for the generation of fairly short output pulses of typically only a few seconds or less. They can produce longer output pulses of up to several minutes, but greater pulse lengths are not really practical as the timing resistor and capacitor values would be impracticable.

Where very long pulse lengths are required it is necessary to use a different type of monostable, and one which is much more complex than conventional types based on devices such as the 555 and 4047. Monostables of this type can be built up from a number of integrated circuits and discrete components, but there is a special precision long timer device, the ZN1034E, which is probably a better solution.

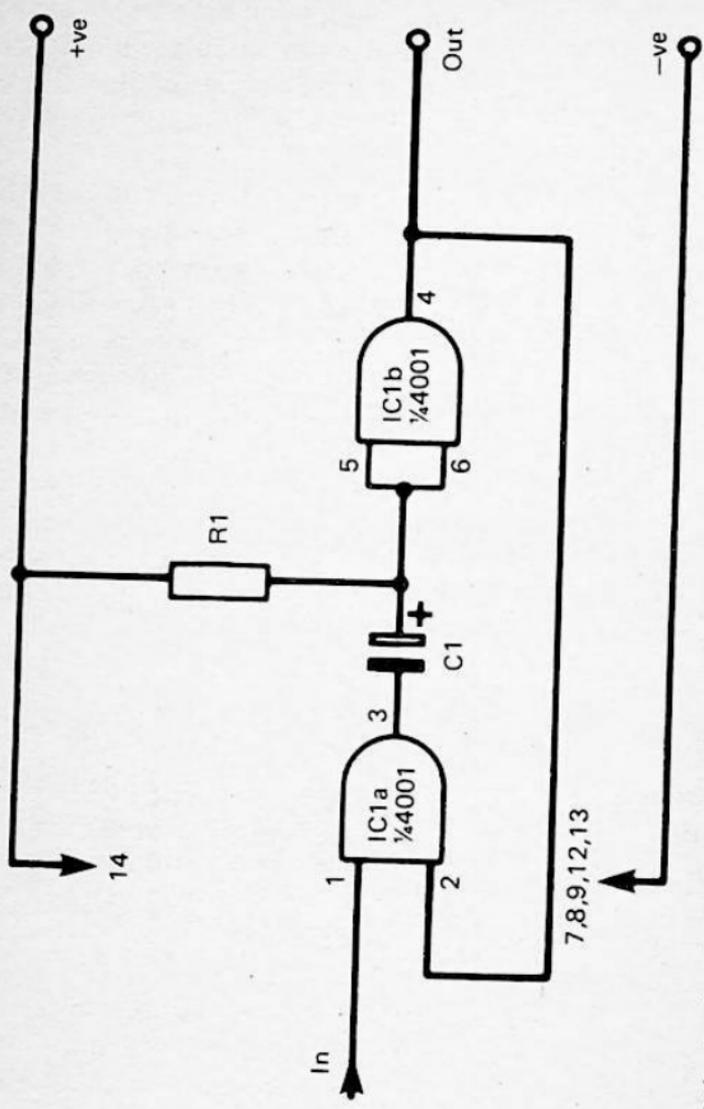


Fig. 36. A simple CMOS monostable circuit

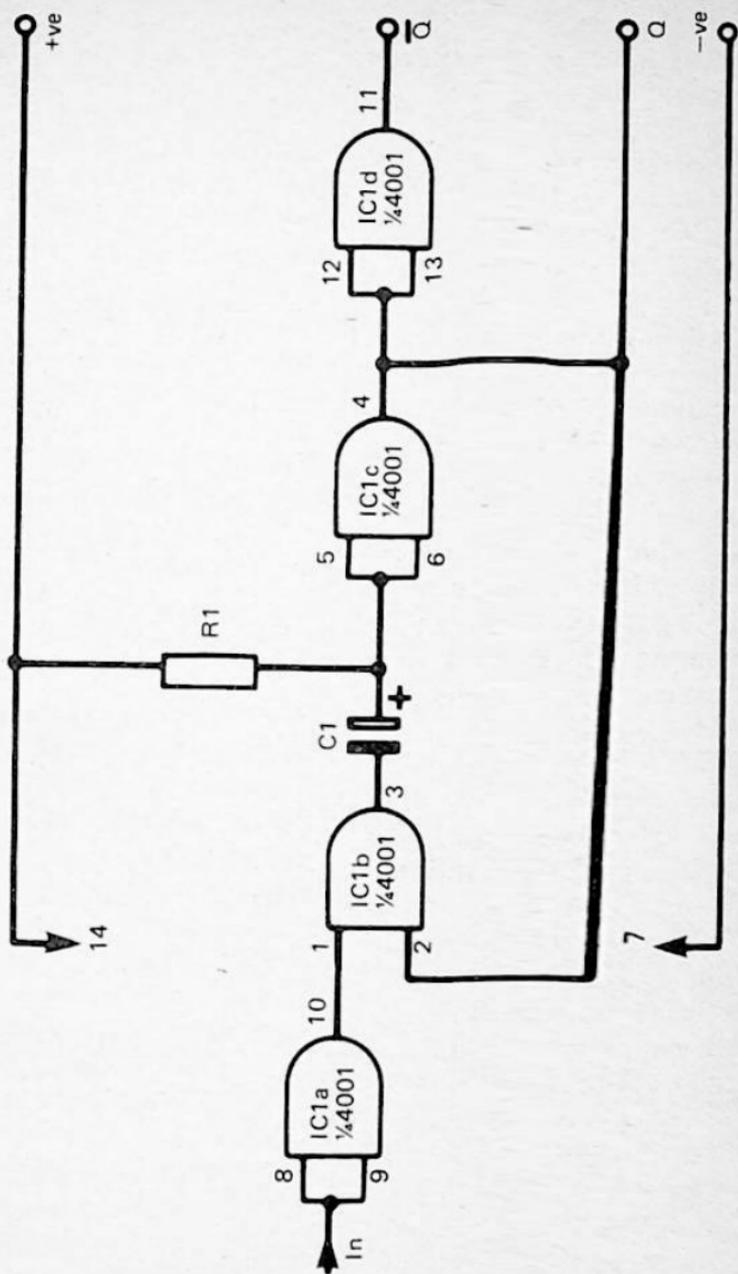


Fig. 37. A negative triggered monostable with Q and \bar{Q} outputs

Precision long timers are basically a C – R oscillator driving a divider chain and logic control circuit, and the output pulse is equal to a certain number of oscillator cycles. The divider chain normally has quite a high total division rate, and the ZN1034E has a 12 stage binary counter. This gives an output pulse which is equal in duration to 4095 oscillator cycles, and this results in a nominal output time of $2736CR$ seconds. This compares to a pulse time of about $1CR$ seconds for most simple monostables ($1.1CR$ seconds for a 555 type for example), and obviously allows a much greater output pulse length to be obtained for a given set of CR values.

Figure 38 shows the circuit diagram of a simple ZN1034E based monostable, and this configuration gives automatic triggering of the circuit at switch-on. Pin 1 is the trigger input, and the ZN1034E is a negative triggered monostable. It is not a retriggerable type and it is quite alright to simply connect pin 1 to earth rather than arranging the circuit so that this pin is supplied with a brief negative pulse at switch-on. Of course, the circuit can be triggered by a negative signal applied to pin 1 if triggering at switch-on is not required, and the trigger input is TTL compatible (as are the outputs). The ZN1034E has both Q and not Q outputs, with a positive pulse being generated at the former and a negative pulse at the latter.

A useful feature of the ZN1034E is an internal 5 volt zener diode which can be used in a shunt regulator circuit to provide the circuit with the necessary stabilised 5 volt supply. R2 is the load resistor for this stabiliser circuit, and the specified value is for a 9 volt supply. This should be changed to 56 ohms for a 6 volt supply, 390 ohms for a 12 volt supply, 560 ohms for a 15 volt supply, and 1k2 for a 24 volt supply. If an external 5 volt stabiliser is used its output is connected to pin 4 of the ZN1034E and pin 5 is left unconnected.

R1 and C1 are the timing components, and as mentioned earlier, the output pulse is nominally $2736CR$ seconds. R1 can have any value between 5k and about 10M, and C1 can have any value over 3n3. It can be a polarised capacitor and the correct polarity is indicated in Figure 38. To give some idea of the long output pulse lengths that can be achieved, timing component values of 2M2 and $100\mu F$ would, for example, give

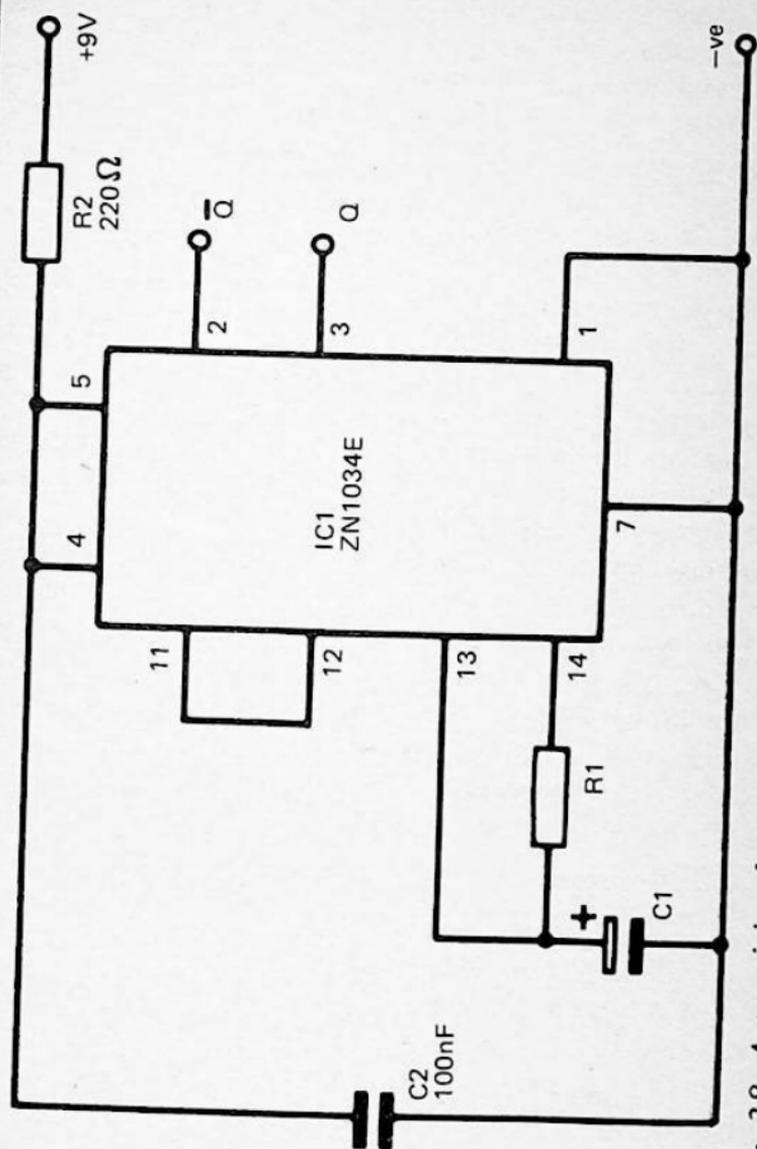


Fig. 38. A precision long timer circuit

an output pulse of approximately one week in duration!

The ZN1034E does not have a reset facility, but this can be achieved by connecting a push to break — release to make switch in series with one lead of C1. When this is operated and C1 is cut out of circuit the small self capacitance of the ZN1034E seems to result in oscillation at a high frequency, and the timing period is brought to an end almost immediately.

VLF Oscillator

It is possible to use the ZN1034E as a very low frequency (VLF) oscillator which provides brief negative pulses from the Q output and complementary positive pulses from the not Q output. This configuration is shown in the circuit diagram of Figure 39.

Here the Q output is low at switch-on and therefore triggers the circuit due to the coupling through R3, and the Q output then goes high. C3 then charges up from the Q output via R3. When the output pulse ends the Q output returns to the low state, but the circuit is not immediately retriggered since C3 must first discharge through R3 and the Q output of IC1. This takes approximately $0.6CR$ seconds, and the maximum practical delay is only a few seconds.

R3 should have a value of about 10k, and C1 can have any value from a few nF upwards. C1 can be a polarised component using the polarity shown in Figure 39.

Calibration Resistor

In the circuit of Figure 38 an internal 100k calibration resistor of the ZN1034E is used by simply connecting pins 11 and 12 of the device together. It is possible to use an external calibration resistor in conjunction with the internal one, and this enables the pulse length to be varied from about $2736CR$ seconds with the discrete resistor at zero, to $4095CR$ seconds with the discrete resistor at maximum value. This simple modification is shown in Figure 40.

It is possible to use an external calibration resistor instead of the internal one, and this enables the pulse length to be varied

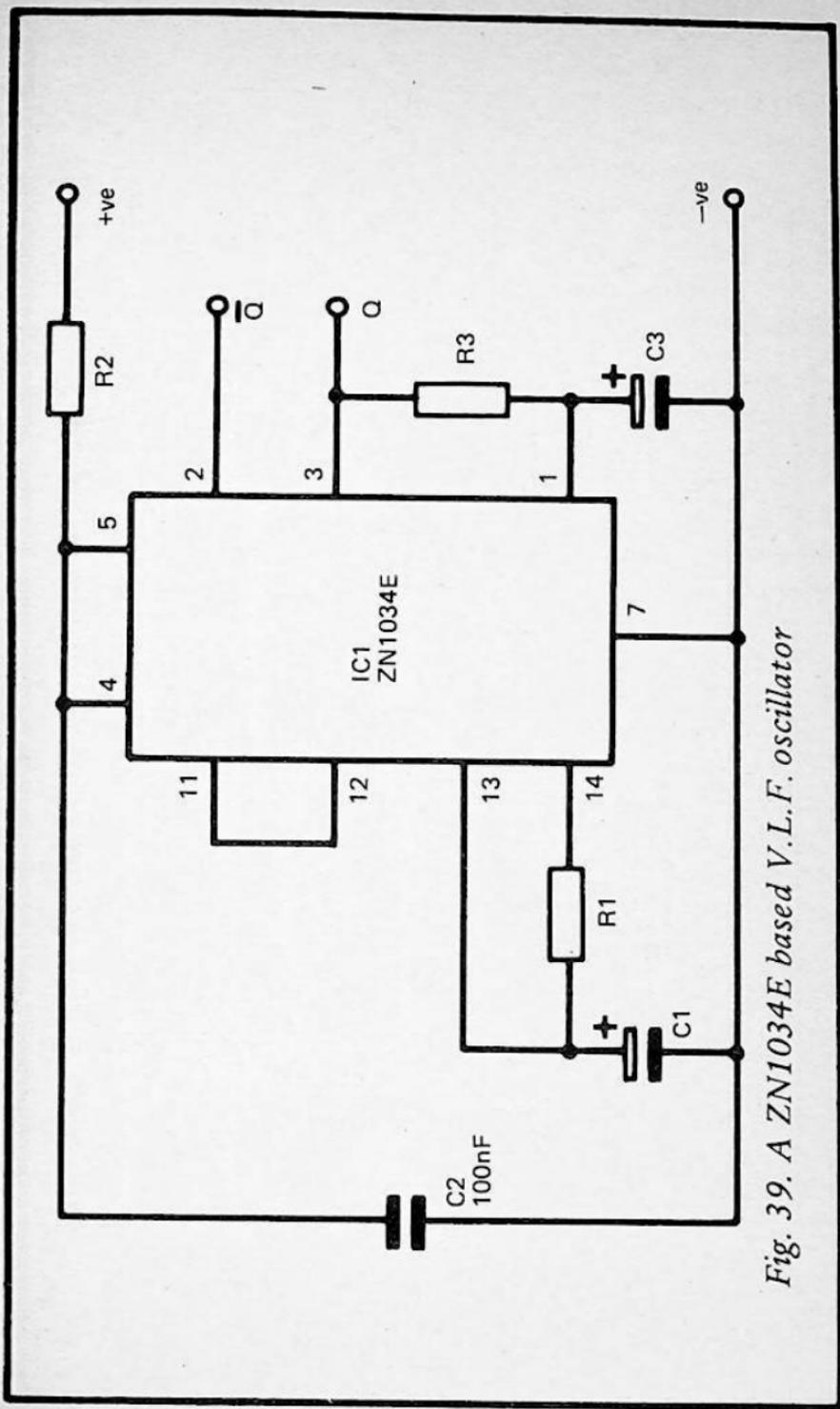


Fig. 39. A ZN1034E based V.L.F. oscillator

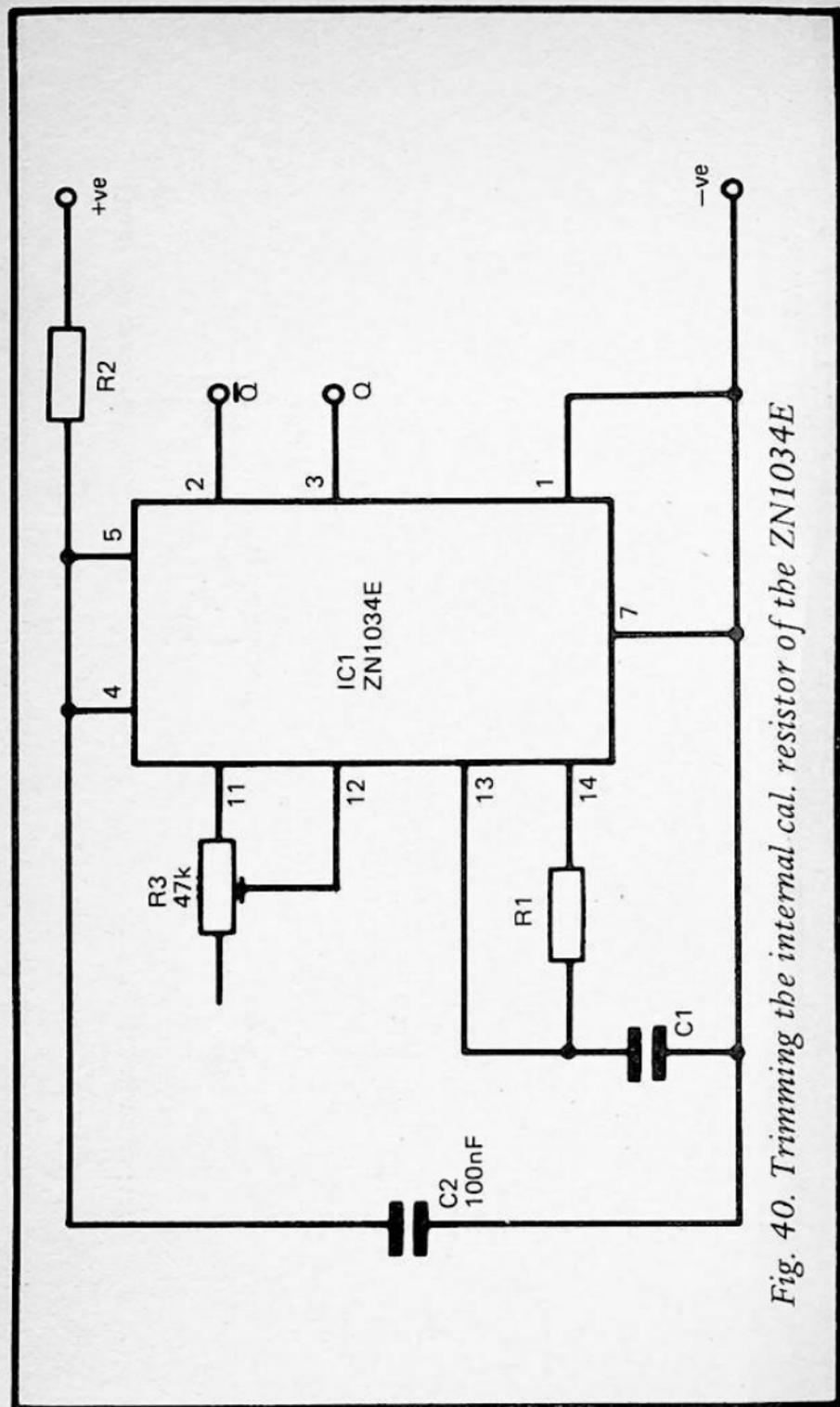


Fig. 40. Trimming the internal cal. resistor of the ZN1034E

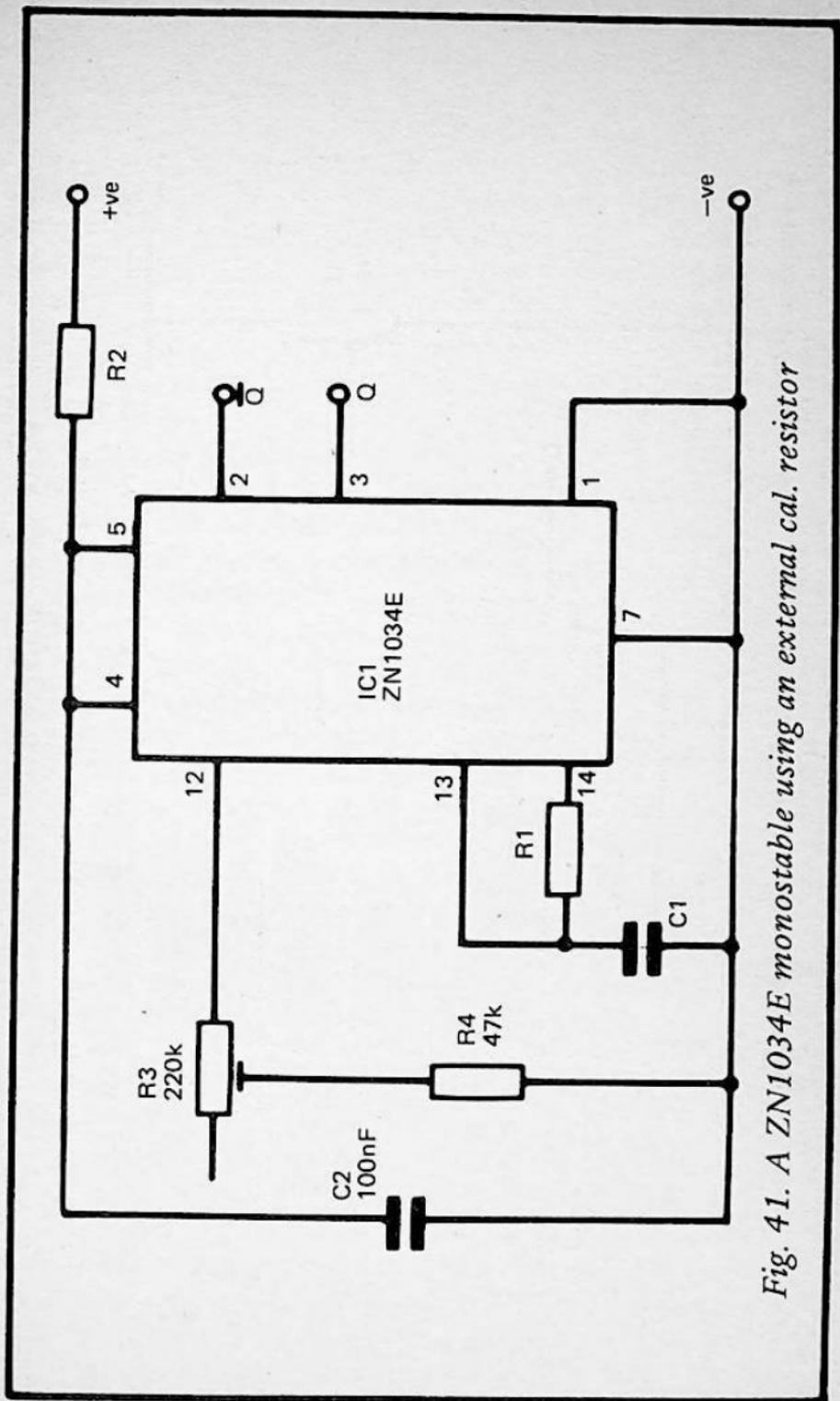


Fig. 41. A ZN1034E monostable using an external cal. resistor

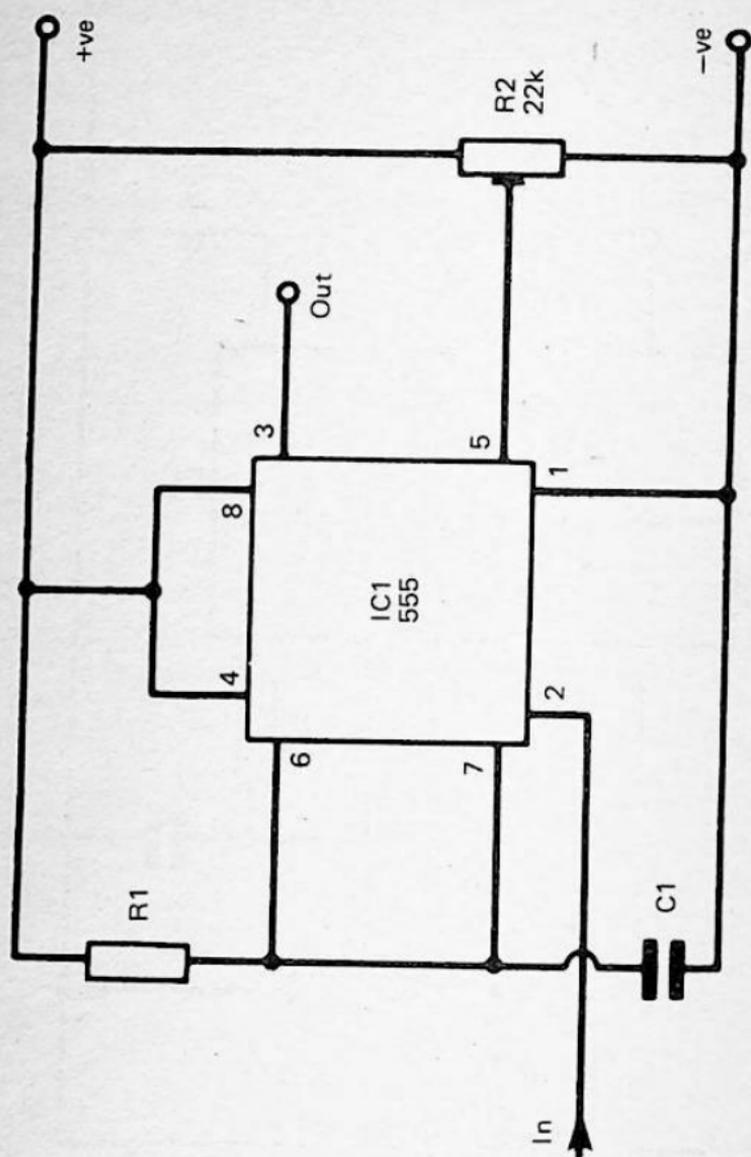


Fig. 42. Trimming a 555 monostable

over the wide range of about 2500CR seconds at minimum value to 7500CR seconds at maximum value. This modification is shown in Figure 41.

This ability to vary the pulse length without altering the main timing component values is extremely useful, particularly in cases where the main timing resistance is a variable type or range of switched resistors having close tolerances. The calibration resistance can then be used to trim the timing range to give the correct limits. The only minor disadvantage is that an external calibration resistor gives slightly inferior stability, but this is not likely to reduce accuracy by a significant amount, and should not normally be of any practical importance.

Trimming the 555

It is also possible to trim the pulse length of a 555 monostable by altering the voltage at pin 5 of the device, and this arrangement is shown in Figure 42. By raising the voltage at pin 5 the C1 threshold voltage at which the timing pulse ends is raised from its normal $2/3 V+$ level, and the output pulse is consequently extended. Reducing the pin 5 voltage reduces this threshold voltage and thus shortens the output pulse duration.

This circuit gives a very wide trimming range of about -85% to +400%.

Chapter 3

MISCELLANEOUS CIRCUITS

Noise Generator

The main use for noise generators is in musical effects units, and cymbal type effects for example are based on a noise signal. In essence a noise signal is very simple, and is just a voltage that changes in a random fashion. It is not therefore possible to show a noise waveform in the same way that a sine or sawtooth waveform can be shown as an oscillograph. If numerous noise oscillographs were taken, due to the random nature of the signal, they would all be different. It is actually difficult to take an oscillograph of a noise signal anyway as a noise signal is comprised of a very wide range of frequencies, and in theory normal ("white noise") consists of all frequencies in equal amounts. When viewed on an oscilloscope a noise signal tends to be little more than a blur.

Generating a low level noise signal is not difficult, and practically any electronic circuit will generate small but significant amounts of noise. Generating a high level noise signal is a little more difficult as it is not a good idea to simply take a minute noise signal and then use a considerable amount of amplification to boost it to the desired level. This can easily result in instability or the breakthrough of mains hum and other noise of the incorrect type!

It is better to use a circuit which generates a reasonably strong noise signal so that only a relatively modest amount of amplification is needed in order to boost the signal to a high level. A circuit of this type is shown in Figure 43, and this circuit is based on zener diode D1. Normally a zener diode is used in a shunt type voltage regulator with a decoupling capacitor fitted across it. This capacitor is needed because the voltage across the zener varies slightly and randomly producing a noise signal that is normally necessary to suppress. In this application though, it is this noise signal that is required and not the stabilised voltage. C2 is therefore used to couple the noise output of D1 to a high gain voltage

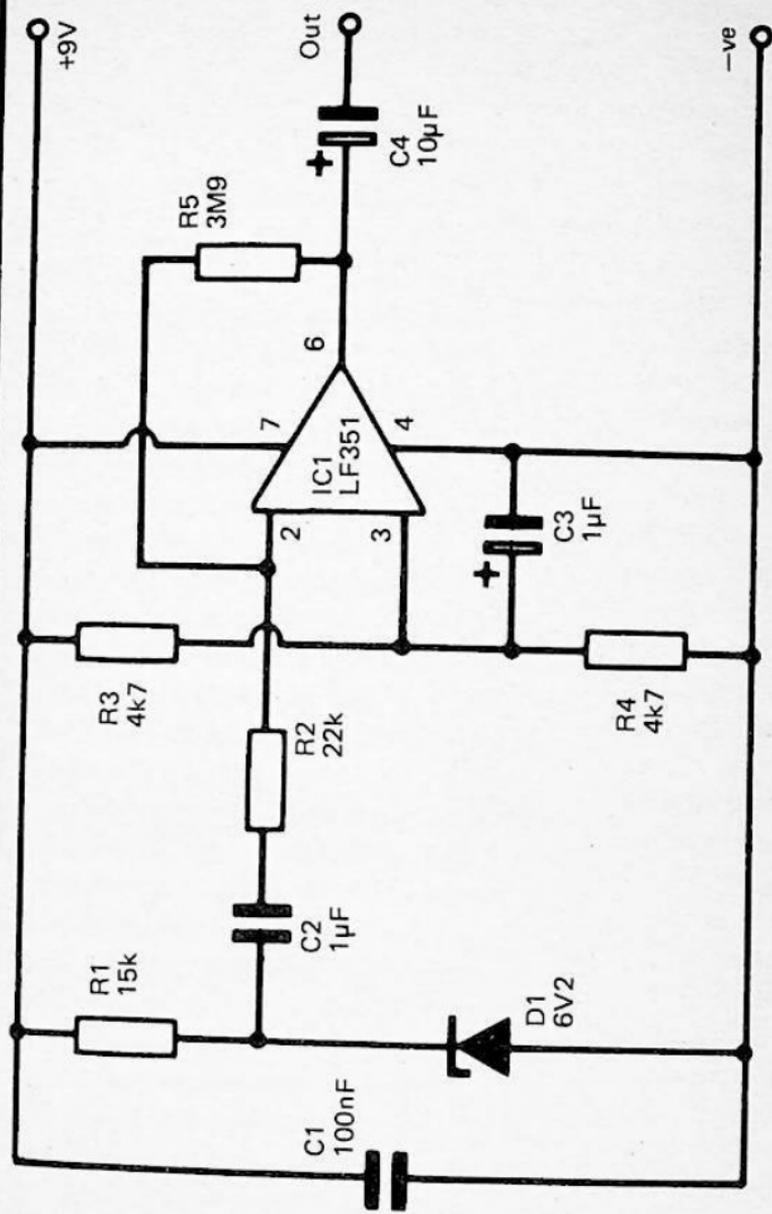


Fig. 43. A simple noise generator circuit

amplifier which utilises IC1 in the inverting mode. This amplifier has a voltage gain of nearly 200 times (46dB) and this boosts the output of the zener diode from approximately 10 millivolts peak to peak to around 2 volts peak to peak. This should give sufficient output for practically any application, and may give too much output for some. In this case the output level can be reduced by lowering the value of R5, and this reduction is proportional to the reduction in R5's value.

Note that the output level will vary somewhat from one zener diode to another, and should therefore be regarded as only very approximate. It is not essential to use a 6V2 zener diode, as specified in Figure 43, and any type having an operating voltage of between about 4V7 and 7V5 should give good results.

There are three alternative noise sources which can be used in this circuit, and these are a germanium diode, a germanium transistor, and a silicon transistor.

A germanium diode is used in exactly the same way as the zener diode, and it is random changes in its reverse leakage current that generate the required noise signal. The leakage current of a germanium transistor can be used if the collector is connected to the negative supply rail, the emitter is connected to the junction of R1 and C2, and the base terminal is left unconnected. This assumes that the transistor is one of the common PNP devices, and the collector and emitter connections must be reversed if one of the few NPN germanium devices is used.

It is unlikely that satisfactory results will be obtained using silicon devices in the manner described above since the leakage currents of silicon devices are usually far too low to give any significant output signal.

A silicon transistor can be used if its base terminal is wired to the negative supply rail and its emitter is connected to the junction of R1 and C2. The collector is left unconnected. The reverse base - emitter breakdown characteristic of the transistor is used as a sort of zener diode, and like a zener diode, generates the required noise signal. Some silicon transistors have breakdown voltages that are too high to permit operation from a 9 volt supply, and it is then either necessary to use a slightly

higher supply voltage or select a suitable device for operation on a 9 volt supply.

When using all three alternative noise sources mentioned above the output level of the circuit is comparable to that obtained using a zener diode, but like a zener diode, the precise signal level obtained will vary significantly from one device to another.

Pink Noise Filter

Some sound effects are based on "pink noise" rather than the white type, and pink noise consists of all frequencies with equal energy in octave bands. In other words, there is the same amount of signal between (say) 100Hz and 200Hz as there is between 1kHz and 2kHz. White noise consists of all frequencies with all frequencies having the same energy level. In other words there is the same amount of signal between 100Hz and 200Hz as there is any other 100Hz wide band, such as 1kHz to 1.1kHz.

Pink noise obviously has a much lower high frequency content than white noise, and it is possible to produce a pink noise signal from a white noise source using a low pass filter. The filter must have a nominal attenuation rate of 3dB per octave, and this cannot be produced using a simple single stage C — R filter as a circuit of this type gives a 6dB per octave roll-off rate.

Instead it is necessary to use a multistage C — R filter circuit such as the one shown in Figure 44. Here each capacitor has a resistor added in series so that the 6dB per octave attenuation rate is never achieved. C1 and R2 operate at the low frequency end of the audio spectrum, R3 and C2 take over at high frequencies, then C3 and R4, and finally C4 at the highest audio frequencies. There is no series resistor for C4 as this does not give a 6dB per octave roll-off rate until the input frequency has gone above the upper limit of the audio range.

IC1 is used as an inverting amplifier which gives about 20dB of gain to compensate for the overall loss of gain through the filter, and it also acts as a buffer amplifier to give a high impedance load to the filter and a reasonably low output impedance.

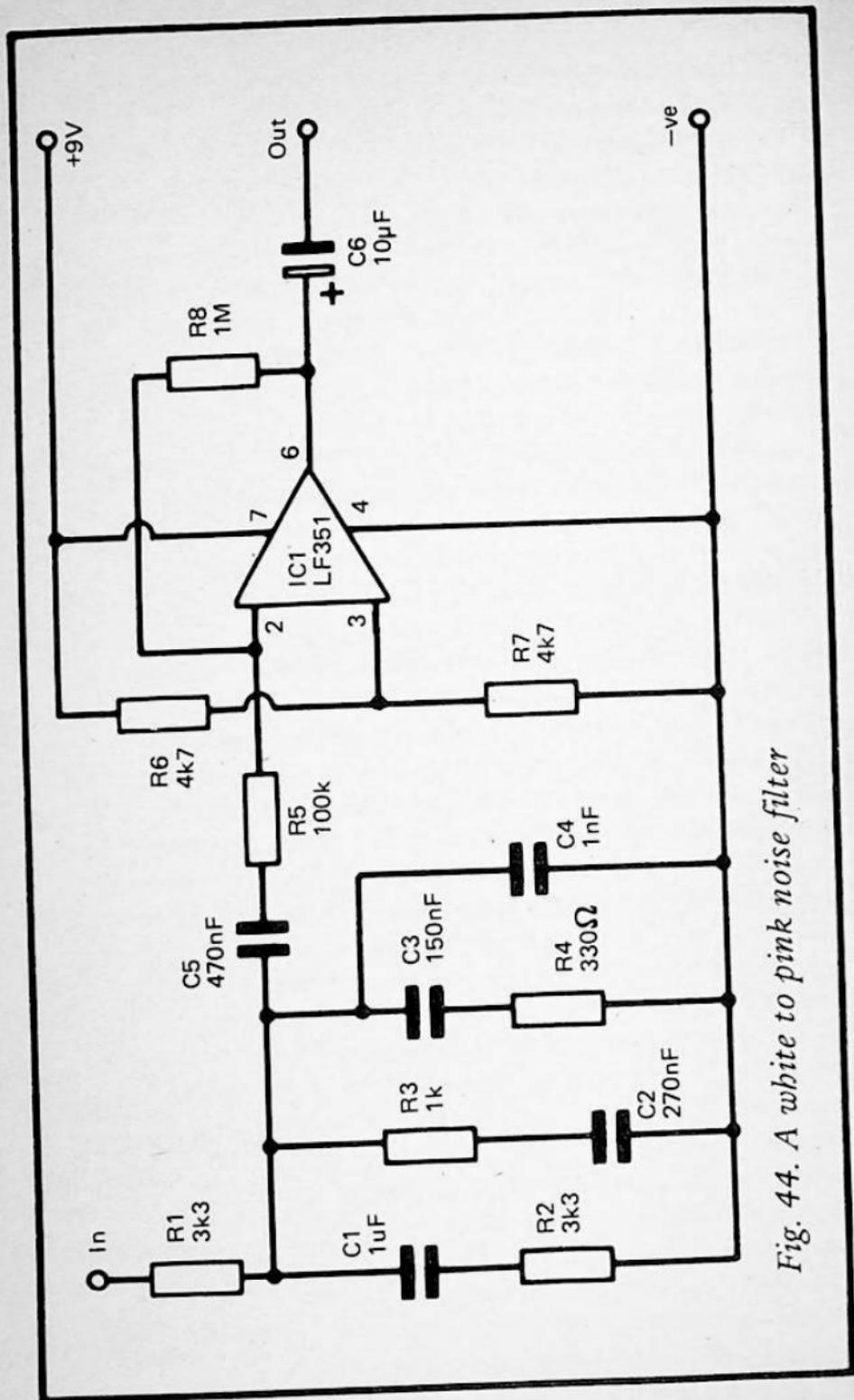


Fig. 44. A white to pink noise filter

Simple Rectifier

It is often necessary to have a circuit that will give a DC output signal that is proportional to the amplitude of an AC input signal. This requires a rectifier and smoothing circuit of some kind, and the simple passive type shown in the circuit diagram of Figure 45 is very useful.

C1 couples the AC input signal to a simple half wave diode rectifier which uses D2, and D1 is included so that current can flow into and out of C1. Without D1 the circuit ceases to function because a charge simply builds up on C1 and it fails to give the required coupling. R1 is the load resistor and C2 is the smoothing capacitor.

The attack time of the circuit is primarily determined by the time constant of C2 and the source impedance of the input signal, and in most applications the source impedance must be fairly low so that a short attack time is obtained. The decay time is determined mainly by the time constant of C2 and R1, and when the input signal ceases the charge potential on C2 falls by 63% after 1 CR seconds. The charge voltage falls to less than 1% of its initial value after 5 CR seconds. An important point to note is that D1 and D2 are germanium diodes and have reverse resistances that may well be no more than a few hundred kilohms each. If R1 is made very high in value it may therefore be found that the decay time is much shorter than expected due to the leakage through D1 and D2. Note also that any load connected across the output will have the same effect unless it has a resistance which is much higher than the value given to R1. The values shown are just given as examples, and would probably need to be changed to suit your particular application. If R1 is made lower in value C1 must be made higher in value in order to maintain the low frequency response of the circuit, and C1 can be made lower in value if R1 is made larger. Any change in the resistance of R1 should be matched by an inversely proportional change in the capacitance of C1. Ideally C1 should not be a polarised capacitor, but it is possible to use a polarised component provided a suitable polarising voltage is present across the component under quiescent conditions and you are careful to connect it with the

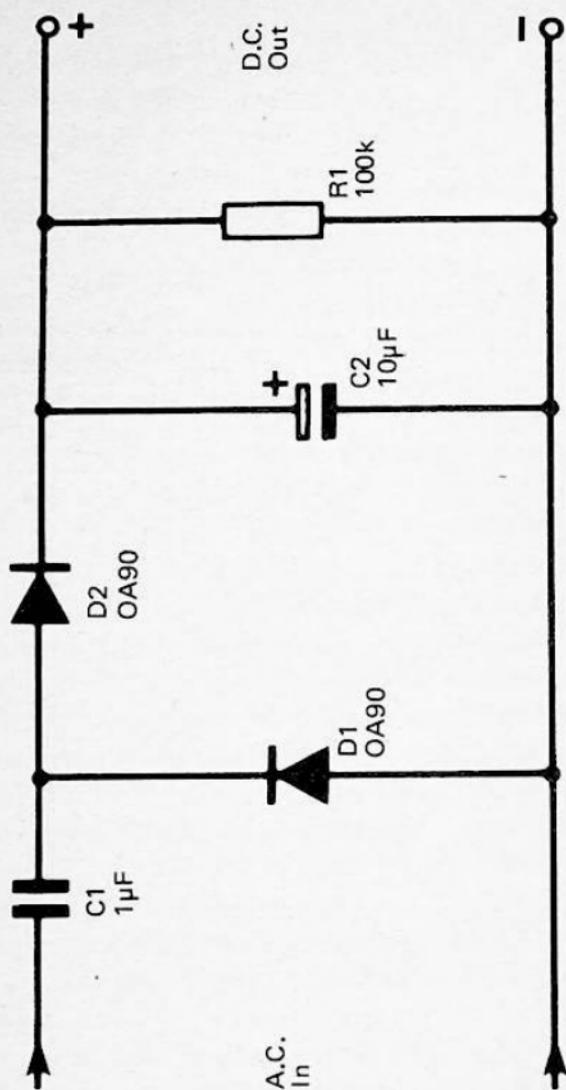


Fig. 45. A simple but useful rectifier and smoothing circuit

correct polarity. In some applications it will be difficult or impossible to calculate the ideal value for C2, but it should be possible to find the optimum value with a little experimentation.

A major drawback of this simple circuit is rather poor linearity, especially at low input levels. This is caused by the forward threshold voltage of D2, and as a germanium diode is employed in the circuit this is likely to be something in the region of 100mV. In theory the DC output voltage should be equal to the peak to peak AC input voltage, but due to the forward voltage drop across D2 (and also that of D1) the output voltage tends to be slightly less than the input voltage. This is obviously most pronounced at low input signal levels where the voltage drop through the diodes is large in comparison to the input voltage, and with very low levels there may be no output signal whatever.

In many applications the linearity of the circuit is not important, but this circuit is obviously unsuitable for use in critical applications such as AC millivoltmeters.

Active Rectifier

In applications where good linearity is required it is necessary to use an active rectifier such as the simple half wave type shown in the circuit diagram of Figure 46.

This is built around an operational amplifier which is powered from dual balanced supplies with a central 0V earth rail. IC1 is used in the non-inverting mode, with the inverting input biased to the 0V rail by R2, and C3 is used to couple the input signal to the non-inverting input.

Negative feedback is provided over IC1 through D1, and at low positive output voltages D1 is not biased into conduction and effectively has a very high resistance. In conjunction with R2 it therefore sets the closed loop voltage gain of IC1 at quite a high figure. On the other hand, when the output of IC1 goes more than about 0.6 volts positive D1 is biased into conduction and effectively provides a resistance of only a few ohms so that IC1 has a closed loop voltage gain of approximately unity.

In practice this means that with a low input voltage IC1 has a high voltage gain so that the output swings about 0.6 volts or

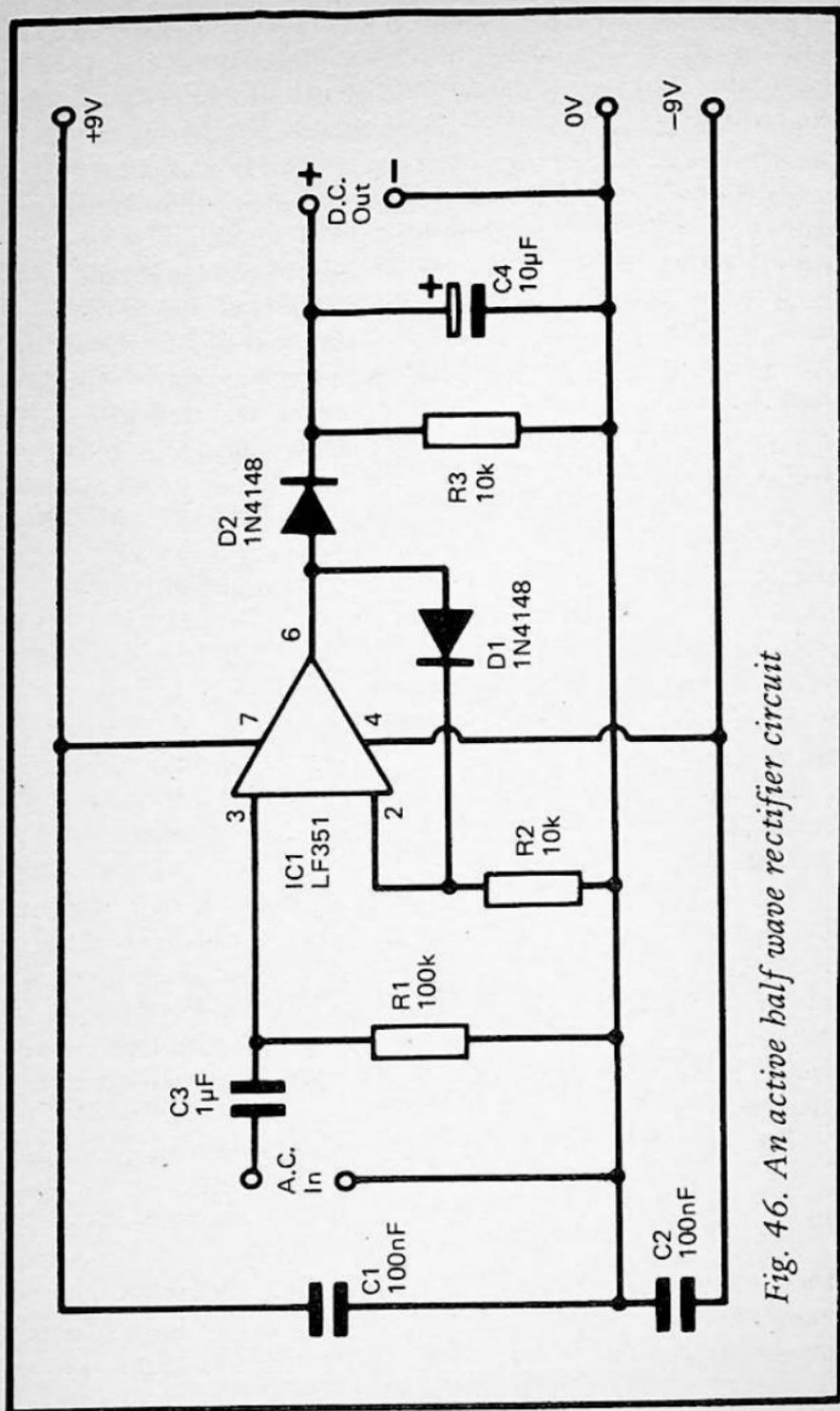


Fig. 46. An active half wave rectifier circuit

so positive almost immediately, and then its gain falls to only about unity so that further rises in input voltage are matched by similar increases in the output voltage. In other words, the output voltage is roughly equal to the input voltage plus about 0.6 volts.

This additional voltage is used to overcome the forward threshold voltage of D2 so that it supplies to R3 and C4 a voltage almost exactly equal to the peak positive input voltage. R3 and C4 are the load resistor and smoothing capacitor, and their values must, of course, be chosen to suit the particular application in which the circuit is used. The attack time is very short since the source impedance provided by IC1 is quite low, but a resistor can be added in series with D2 to give a longer attack time if necessary.

D1 and D2 are both silicon diodes, and in this type of circuit these have the advantage that the reverse resistance is very high and R3 can therefore be made very high in value if desired without the reverse resistance of D2 having any significant shunting effect on it. Germanium diodes such as the OA90 and OA91 will work in this circuit, and due to their lower forward threshold voltages will probably give slightly better linearity, especially at high frequencies. However, the linearity of the circuit using silicon diodes is more than adequate for any normal application, and the circuit works well at frequencies as high as 200kHz.

The maximum output that can be produced is about one volt less than the positive supply rail potential, or about 8 volts using the specified 9V - 0V - 9V supply. The circuit can use a supply voltage of up to about plus and minus 15 volts (plus and minus 18 volts is the maximum for the LF351 and similar operational amplifier devices), and the maximum output can thus be boosted somewhat if required.

The current consumption of the circuit is only about 2mA.

PSU Rectifiers

There are only two types of rectifier commonly used in power supply circuits; the push-pull type and the bridge type. These are shown in the circuit diagrams of Figure 47 and 48 respectively.

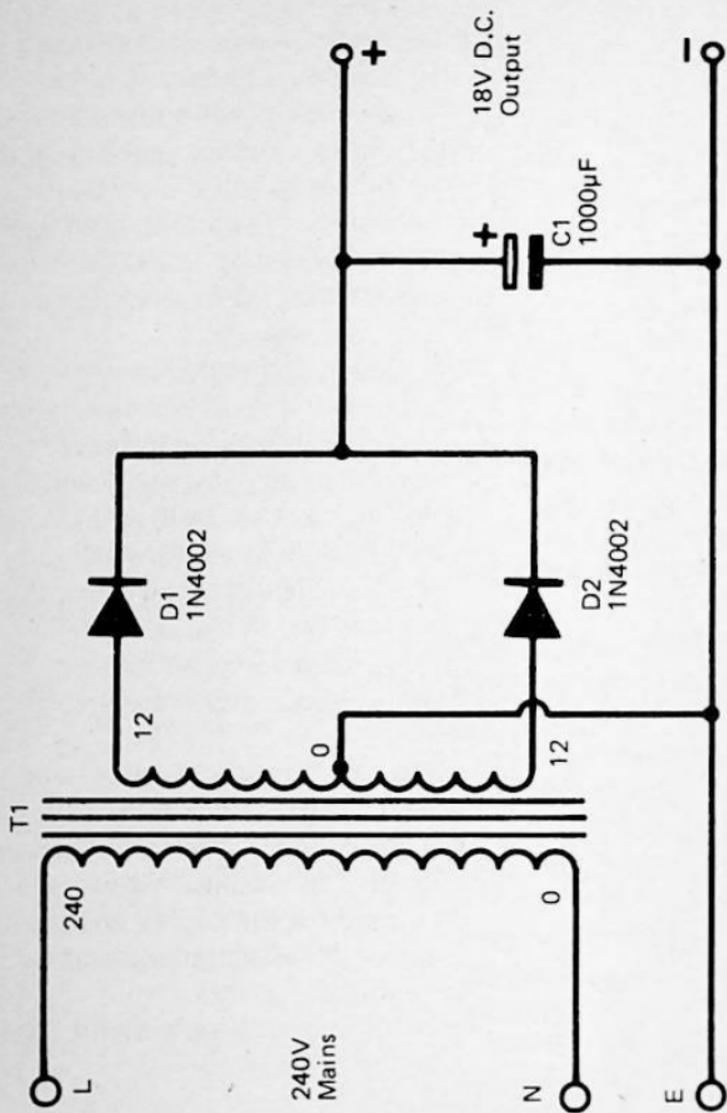


Fig. 47. A power supply using push-pull rectification

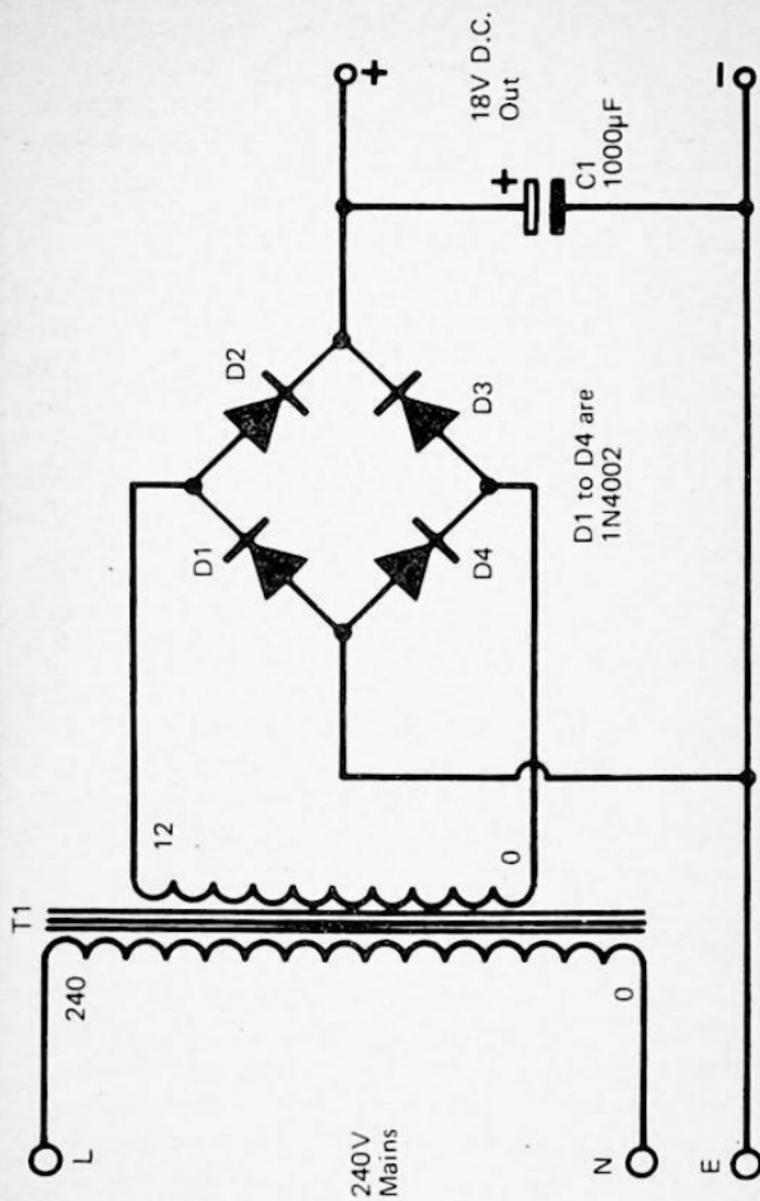


Fig. 48. A simple bridge rectifier power supply

Both are full wave types, and half wave circuits are only rarely used in power supplies because of their much lower efficiency compared to full wave types.

The push-pull type is very simple in operation, and requires only two rectifiers, but needs either a centre tapped transformer or one having two identical secondaries connected in series to effectively produce a centre tap. The secondary winding of step-down and isolation transformer T1 provides out-of-phase signals to D1 and D2 so that D1 supplies pulses of current to C1 on one set of half cycles, while D2 supplies pulses of current to C1 on half cycles of the opposite polarity.

A point which is often overlooked, but which is crucial to power supply design, is that the rectifiers only supply current to the smoothing capacitor during the brief periods when the transformer is supplying a voltage which is about 0.5 volts or more than the charge potential on C1. Thus, although there may be an average output current of (say) one amp from the supply, a current of perhaps ten amps or more may be passed through the rectifiers during the brief charge periods.

This is not really relevant when it comes to choosing rectifiers with an adequate rating, since the current rating normally quoted for rectifiers in brief data is an average or RMS figure. The peak current rating is usually very much larger than this figure (some data actually states the maximum smoothing capacitor value that can be used with the component concerned). What it is important to bear in mind is that the transformer will be heavily loaded during these brief current pulses and may give a much lower voltage than one might expect.

The voltage rating of a transformer is normally the RMS output voltage at full load, and under low or zero load conditions the output voltage will be somewhat higher than this. The smoothing capacitor charges to the peak output voltage (minus a voltage drop of about 0.6 volts or so through the rectifiers) of the transformer, and this is 1.41 times the RMS voltage. Thus, although a 12 volt transformer is shown in Figure 47, the unloaded DC output voltage is about 18 volts, and could actually be as high as 20 volts or so. This voltage drops substantially as the loading on the supply is increased towards maximum, and typically an output voltage of a little

more than the voltage rating of the transformer is obtained at full load. However, this depends upon how conservatively or otherwise the transformer used in the circuit is rated, and using transformers from a variety of manufacturers, but all having the same secondary ratings, would probably give a range of significantly different loaded and unloaded output voltages.

One reason for this is that a transformer having a fairly high unloaded output voltage and a high secondary impedance would give the same loaded voltage as one having a comparatively low secondary voltage and a corresponding lower secondary impedance. With a resistive load placing full load on the transformers they would give identical results, but not under any other load conditions. Another possibility is that the manufacturers ratings simply apply to slightly different load conditions.

In practice this makes it difficult to predict exactly what output voltage will be produced at various output currents, and one can only make an estimate unless full data on the transformer is available, or you are using a component which you have used before and which you are therefore familiar with. If it is not possible to select a transformer of known performance it is probably best to choose one having a secondary current rating a little higher (say about 20% or more) than the output current required to ensure that there is sufficient output voltage available at full load.

The bridge rectifier has the advantage of needing only a single, untapped secondary winding, but an important point to keep in mind is that the maximum continuous output current that can be obtained using this configuration is approximately 0.62 of the secondary current rating, and this is due to pulsed rather than continuous loading on the transformer. This figure is affected to some extent by the size of the smoothing capacitor used, with a large amount of smoothing giving a greater derating factor as this gives shorter, heavier current pulses into the smoothing capacitor. The characteristics of the transformer also have some effect on the derating factor, but when using a bridge rectifier it is advisable to choose a transformer having a current rating at least 50% higher than the maximum output current unless maximum loading will only be brief and intermittent.

No derating is needed when using a push-pull rectifier.

The component values shown in Figure 47 and 48 are suitable for output currents of up to about 1 amp and with maximum unloaded output voltages of up to about 48 volts. Higher currents would require the rectifiers to be replaced with higher current types, such as 1N5401 rectifiers which can handle maximum average currents of up to 3 amps. Higher voltages would require rectifiers having suitably higher PIV (peak inverse voltage) ratings, and the 1N4003 (1 amp) and 1N5402 (3 amp) devices have a PIV rating of 200 volts. As the peak voltage across each rectifier can be as much as just over double the output voltage of the supply, 200 volt rectifiers are suitable for output voltages of up to about 98 volts.

The smoothing capacitor should have a value of at least $1\mu\text{F}$ per mA of output current, and an output of 2 amps for example, would need a smoothing capacitor of at least $2200\mu\text{F}$ (the nearest preferred value to $2000\mu\text{F}$), and a higher value such as $4700\mu\text{F}$ would be better.

With any of the four rectifier circuits described in this book the polarity of the output signal can be reversed simply by reversing the polarity of the diodes or rectifiers (and the smoothing capacitor must also be connected with the opposite polarity, of course).

If an on/off switch is required this is fitted in the primary circuit of T1 and should be a double pole type so that T1 is completely cut off from the mains supply when the unit is turned off. As with any circuit that is connected to the mains supply, the normal safety precautions should be observed to ensure that there is no risk of anyone receiving an electric shock from the unit.

Dual Supplies

Some circuits, especially those which are based on operational amplifiers, require dual balanced power supplies. The usual arrangement used to provide dual balanced supplies is shown in Figure 49, and this is really just two push-pull rectifier and smoothing circuits powered from a common mains transformer, one circuit producing the positive supply and the other giving

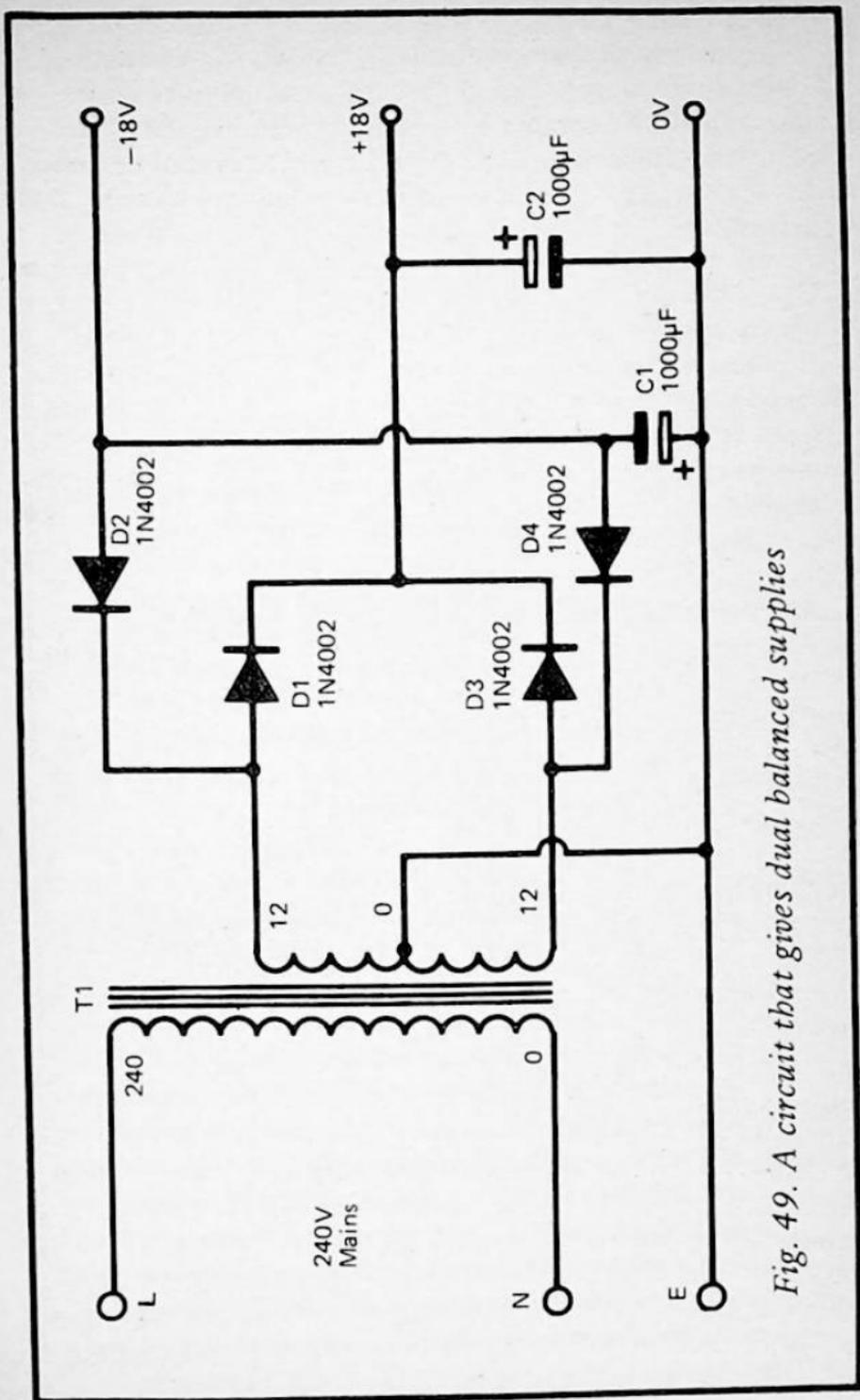


Fig. 49. A circuit that gives dual balanced supplies

the negative rail.

The loading on the transformer is higher than when using a single push-pull rectifier, and accordingly the secondary current should be derated by a nominal factor of 0.62. Of course, the circuit values shown in Figure 49 are only examples, and must be altered to suit your requirements.

Series Regulator

In many applications a simple unregulated power supply is unsuitable due to the large variations in output voltage that occur with changes in loading, and to a lesser extent changes in output voltage caused by variations in the mains supply voltage. Another important factor is the ripple on the output of a non-stabilised supply. At high output currents this is likely to be at least a few hundred millivolts RMS even if a high value smoothing capacitor is used, and this is far too high for the majority of applications.

Using a simple stabilised supply the output voltage can be maintained to within about 1% of its normal level, and hum and other noise on the output can be reduced to only a millivolt or so. Basically a regulated supply is a circuit of the type just discussed in the previous sections of this book, but with a voltage regulator added at the output. The step-down, rectifier, and smoothing stages are exactly the same for regulated and non-regulated supplies. Here we will therefore only consider the voltage regulator section of stabilised supplies.

Probably the most simple way of obtaining a stabilised supply in the majority of applications is to use one of the three terminal voltage regulator devices that are available at very attractive prices. These use the simple method of connection shown in Figure 50, and the only discrete components required are two decoupling capacitors which are needed to aid stability and the transient response of the circuit. The 78-- series of voltage regulators are available in a number of voltages, and the nominal output voltage is indicated by the last two digits of the type number. For example, a 7805 is a 5 volt regulator, a 7812 is a 12 volt type, and so on. This series of integrated circuits can handle output currents of up to 1 amp, and they

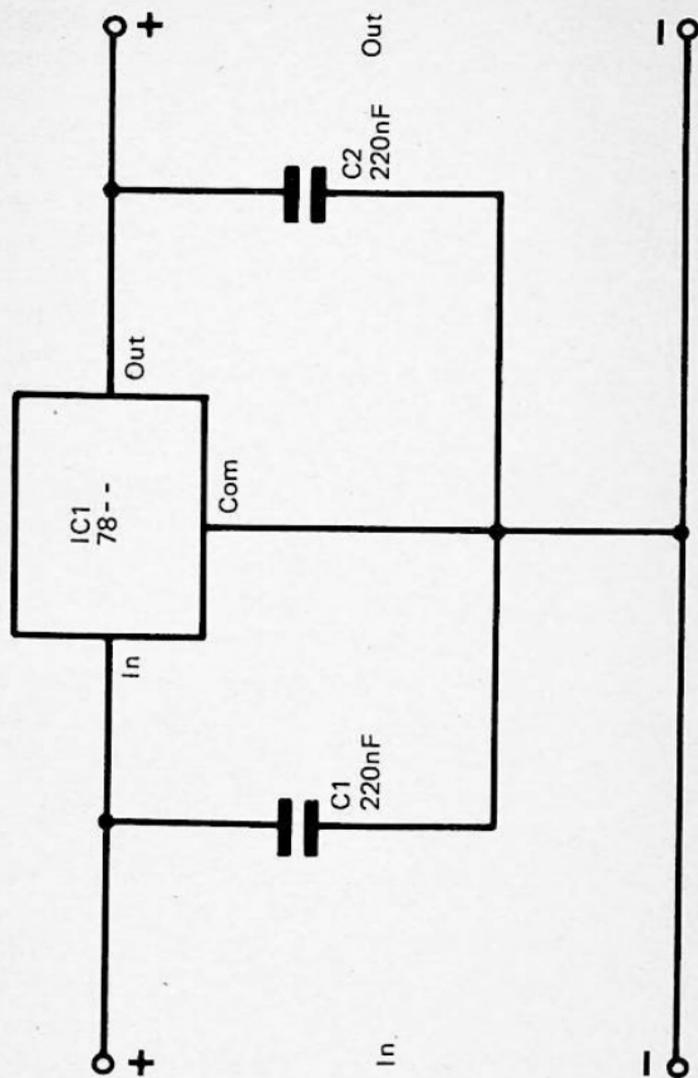


Fig. 50. A simple series regulator circuit

have output current limiting which prevents the device being damaged by a short circuit or overload at the output. They also have thermal shut-down protection circuitry.

An important point to have in mind when designing a stabilised power supply using these devices is that the input voltage must never fall to less than about 2.5 volts more than the output potential. The circuit will not function properly if the input voltage is allowed to fall below this level, giving a fall in output voltage and an increase in the ripple level. In order to take account of the fairly high ripple level that is likely to be present on the input to the regulator, and to give some margin for error, it is advisable to have an input voltage that is about 3.5 to 4 volts above the nominal output voltage under maximum load.

There is a limit to the maximum input voltage that these regulators can withstand without sustaining damage, and for most devices this is well above the output voltage at something in the region of 35 volts (the exact figure varies somewhat from one device to another, and from one manufacturer to another).

When used at high output currents there is going to be a dissipation of around 3 to 6 watts in the regulator, and this makes it necessary to fit the device with a reasonably large heatsink to ensure that overheating and shut-down do not occur.

There are a number of variations on the basic 78— series of integrated circuits, and one of these is the 78L— series of regulators. These are low power versions of the 78— regulators and can provide a maximum output current of 100mA. They are in a plastic TO-92 style case and cannot be fitted with an effective heatsink, and it is therefore necessary not to exceed the maximum dissipation figure of 500mW. The 78M— series of integrated circuits are slightly lower power versions of the 78— series having a maximum output current of 500mA.

For each device in the various 78— series of regulators there is a complementary 79— series integrated circuit. Whereas the 78— devices give a positive output the 79— series of regulators give a negative output. They are used in the same circuit except for the reversal of the polarity of the input voltage (and the consequential change in polarity of the output signal

that this gives). The 79— series of integrated circuits are obviously ideal where dual stabilised supplies are needed, and would then be used in conjunction with the complementary 78— series integrated circuit.

The two decoupling capacitors should ideally be placed very close to the regulator in order to minimise the risk of instability, and this is perhaps a little more important with 79— series devices than with 78— series types.

Boosted Voltage

Although a number of output voltages are available using three terminal voltage regulators, from time to time a supply voltage for which a suitable regulator is not available may be required. In some cases it is possible to boost the output of a three terminal regulator to the required supply potential, and a simple way of doing this is shown in Figure 51.

The output terminal of a three terminal regulator is stabilised at a certain level above the potential at the common terminal of the device (which is normally at 0V, of course). By raising the voltage at the common terminal it is possible to boost the output voltage, and in this case the output voltage is boosted by a little over 1 volt by the simple shunt regulator formed by R1, D1, and D2. D1 and D2 are silicon rectifiers that are forward biased by R1, and their forward threshold voltage is used to give a sort of zener diode avalanche effect and thus give a stabilised voltage. The output voltage is boosted by approximately 0.6 volts per rectifier, or by about 1.2 volts using the circuit of Figure 51. However, the output can be boosted by about 0.6 volts using a single diode, or 1.8 volts using three diodes in series.

The protection circuitry of the regulator should remain effective provided no more than three diodes are used in the booster circuit, and the performance of the circuit will not be significantly degraded in comparison to that obtained using the same regulator in a non-boosted circuit.

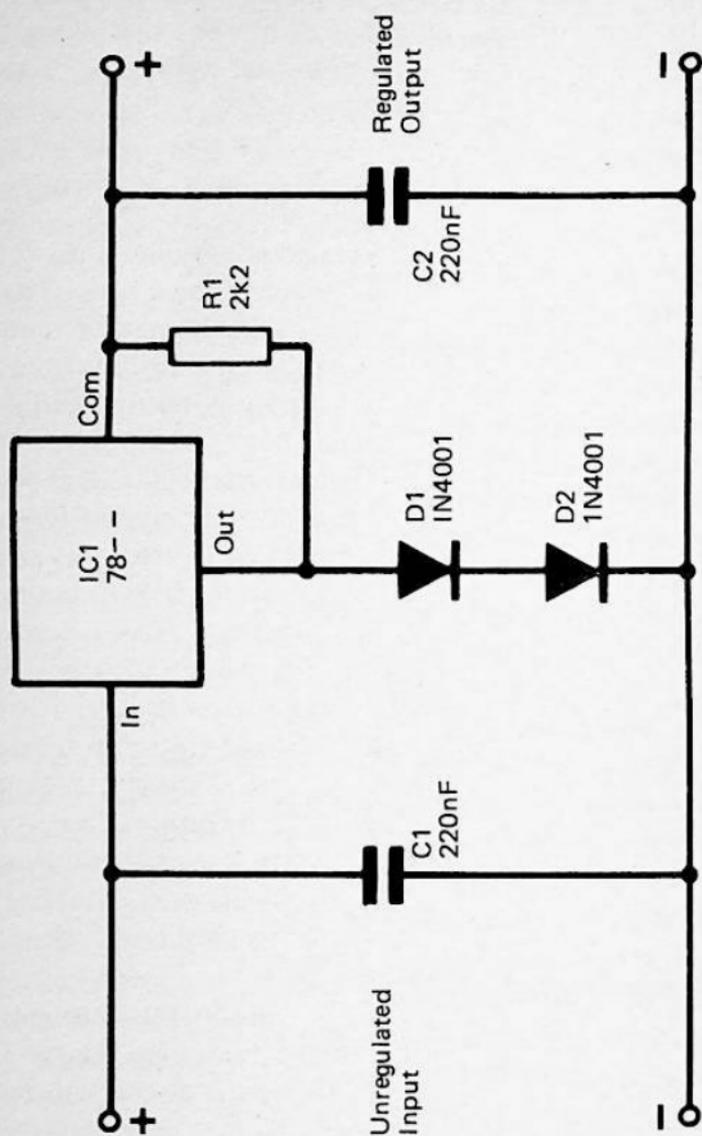


Fig. 51. Boosting the output voltage of a three terminal regulator

Variable Regulators

In some applications it is necessary to have a supply voltage that cannot be achieved using a straight forward three terminal regulator circuit or a boosted type, and in such cases the 78GU1C 1 amp variable voltage regulator integrated circuit (or the 78MGU1C 500mA version) can be used. These are four terminal devices, and the additional terminal is used in conjunction with a potential divider network to set the output voltage at any desired figure between 5 and 30 volts, as shown in Figure 52.

The voltage at the control (cont.) terminal of the device is stabilised at 5 volts by a negative feedback action from the output of the circuit. Thus if R1 has a value of zero ohms the output will be stabilised at 5 volts, like the control terminal. If R1 has a higher value there will be a voltage drop across this component and the output will be equal to 5 volts plus this voltage drop. The output voltage can be calculated from the following formula:-

$$V_{out} = [(R1 + R2)/R2] \times 5 \text{ volts}$$

For example, if R1 and R2 are both 4k7, this gives $4k7 + 4k7 = 9k4$, $9k4$ divided by $4k7 = 2$, and $2 \times 5 = 10$ volts. The current through R1 and R2 should be about 1mA, which means that the total resistance through R1 and R2 should be roughly equal in kilohms to the output potential in volts (e.g. about 10k for an output of 10 volts).

In order to obtain reasonably accurate results R1 and R2 should be close tolerance (2% or better) resistors, or a preset potentiometer could be used instead so that the output voltage can be trimmed to the desired figure. An ordinary potentiometer can be used if a variable voltage workshop power supply is required.

The maximum permissible input voltage for the 78GU1C and 78MGU1C integrated circuits is 40 volts. Negative versions of these devices, the 79GU1C and 79MGU1C are available, incidentally.

L200 Regulator

Probably the most versatile voltage regulator integrated circuit

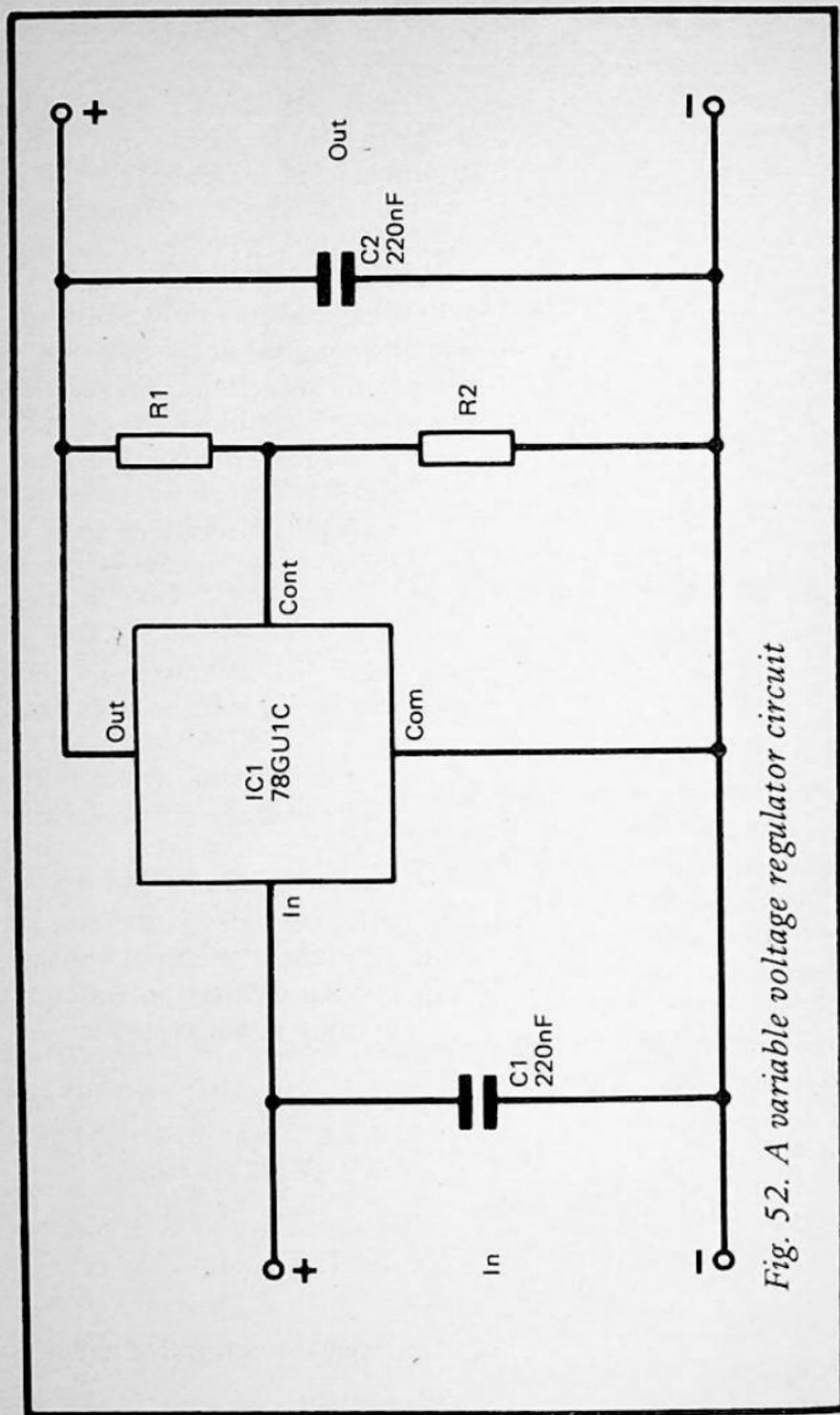


Fig. 52. A variable voltage regulator circuit

available at present is the L200 which has an output voltage range of 2.85 to 36 volts and can handle output currents of up to 2 amps. An important feature of this device is the ability to adjust the output current at which the current limiting comes into effect. With other devices this is preset at the maximum figure the device is designed to handle, and it may well be substantially higher than the maximum output current which the power supply as a whole is designed to provide. With the L200 the output current can be limited to a level that matches the capabilities of the rest of the circuit. Figure 53 shows the basic circuit for the L200 regulator.

The control terminal of the L200 (pin 4) is stabilised at a nominal potential of 2.77 volts, and this enables the output voltage to be taken to a lower level than is possible with the 78— and 79— series adjustable regulators discussed earlier. The formula for determining the output voltage is essentially the same as for the 78GUIC, etc., but must be modified slightly to take into account the lower voltage at the control terminal. The modified formula is as follows:-

$$V \text{ out} = [(R1 + R2)/R2] \times 2.77 \text{ volts}$$

The current through R1 and R2 should be about 3mA, which means that R2 should have a resistance of between about 680 ohms and 1 kilohm.

Note that the input voltage should not exceed 40 volts.

R3 sets the maximum output current, and the value of this resistor is calculated using the following formula:-

$$R3 = 0.45/I \text{ out max ohms}$$

For example, a maximum output current of 0.8 amps would require a resistor value of 0.45 divided by 0.8 = 0.5625 ohms. The nearest preferred value below this figure is 0.56 ohms, and this is the value that would be used. Do not use the nearest preferred value above the calculated figure as this would limit the output current to less than the required figure. Resistors of less than 1 ohm only seem to be available in a limited range of values, and in practice it might be necessary to use a 0.47 ohm component or make up the required value from two or three resistors in series or parallel. For instance, two 1 ohm resistors

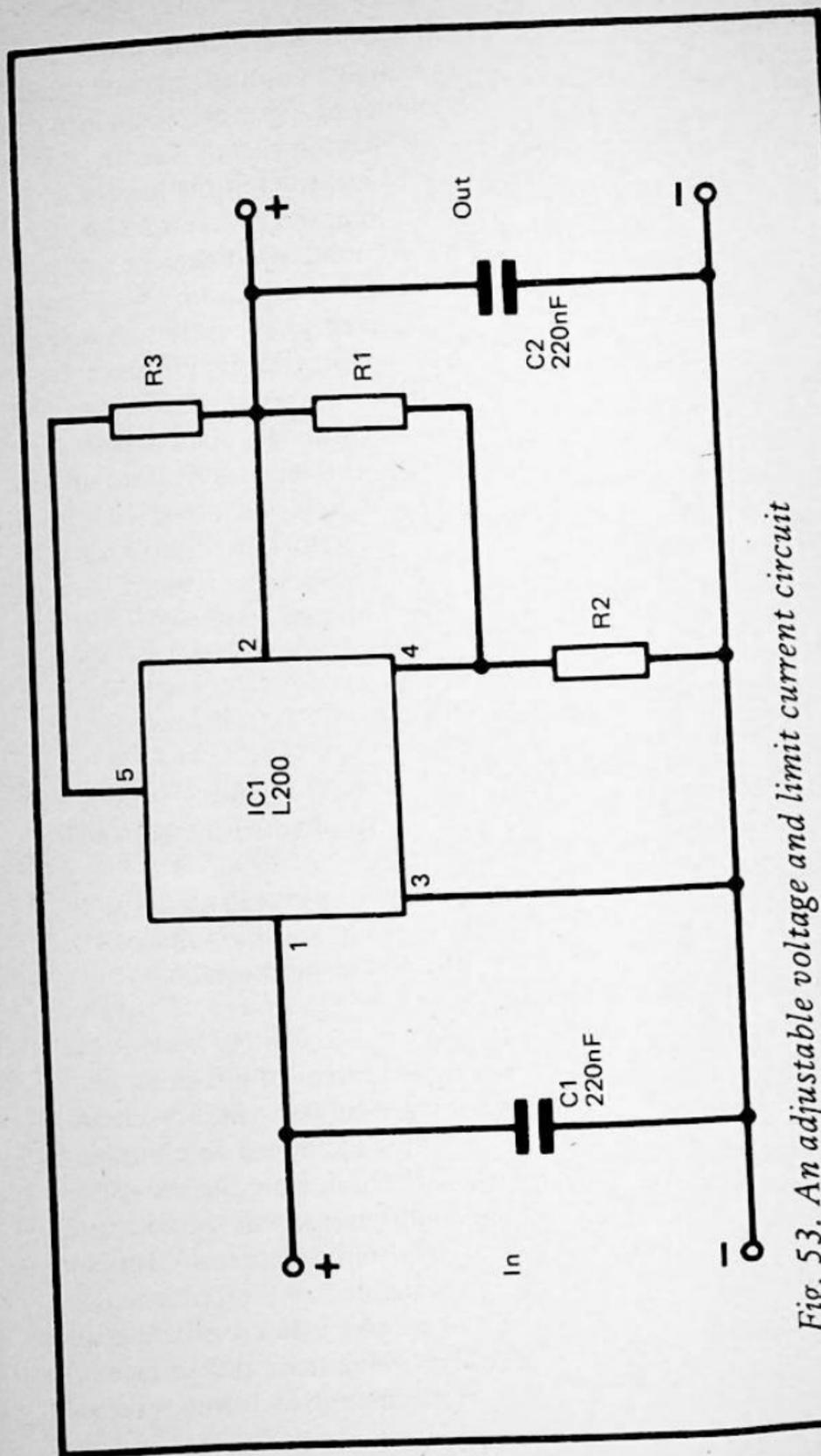


Fig. 53. An adjustable voltage and limit current circuit

in parallel would give a value of 0.5 ohms which is close enough to the calculated figure in the example given above. A 1.2 ohm and 1 ohm component in parallel would give better accuracy with a combined resistance of 0.5454 ohms, but due to component tolerances this might limit the output current to a level just below the required level, and in practice two 1 ohm resistors in parallel or a single 0.47 ohm type might be a better choice.

For output currents of up to 0.5 amps R3 should have a power rating of at least 0.25 watts, and a rating of at least 0.5 watts for output currents of between 0.5 amps and 1 amp. The rating should be at least 1 watt for output currents of between 1 amp and the maximum permissible output current of 2 amps.

Simple Series Regulator

The simple series regulator circuit of Figure 54 is useful for output currents of up to about 50mA. The circuit obviously does not incorporate thermal or output overload protection circuits and care must be taken not to damage Tr1 due to excessive current flow or overheating (the P_{tot} rating for the BC109C is 300mW).

The circuit is just a simple zener shunt stabiliser using D1, R1 and C1, plus an emitter follower buffer stage using Tr1, with the output of the shunt stabiliser feeding into the input of Tr1. This gives a stabilised output at the emitter of Tr1 that is about 0.7 volts below the zener voltage, this reduction in voltage being due to the base - emitter stand-off voltage of Tr1. Thus D1 should have an operating voltage that is about 0.7 volts above the required output potential. Alternatively a silicon diode such as a 1N4148 can be connected in series with D1, but with the opposite polarity so that it is forward biased and boosts the input voltage to Tr1 by about 0.7 volts. This counteracts the voltage drop through Tr1 and gives an output voltage that is approximately equal to the operating voltage of the zener diode used in the D1 position.

An advantage of this circuit over most other series regulators is that it gives quite good regulation until the input voltage falls to only a few hundred millivolts more than the zener voltage,

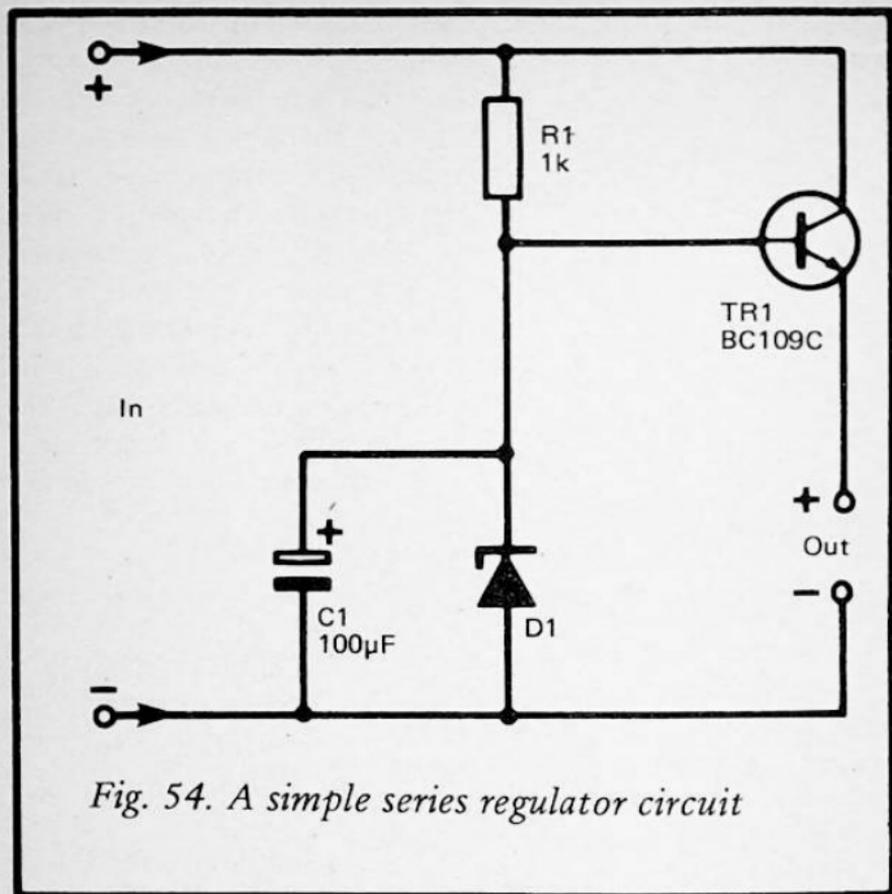


Fig. 54. A simple series regulator circuit

and it will therefore work quite well with the input voltage only about 1 volt higher than the output voltage.

The current consumption of the circuit is largely dependent on the input voltage, and is approximately 1mA for each volt that the input voltage is higher than the operating voltage of D1. Thus the current consumption would typically only be about 2 or 3mA.

Constant Current Generator

It is sometimes necessary to have a circuit which will give a constant current flow through a load resistance that will vary considerably. We have already seen an example of this in the linear sawtooth oscillator circuit which was described earlier.

Figure 55 shows the circuit diagram of a simple but efficient constant current generator. R2 plus D1 to D3 form a simple shunt type voltage regulator circuit having an output voltage of about 2.1 volts, and this voltage is used to bias the base terminal of Tr1. The voltage at the emitter of Tr1 is about 0.7 volts less than the base potential, or about 1.4 volts in other words. This voltage, plus the value given to R1, sets the emitter current of Tr1.

The collector current of Tr1 will be very much the same as the emitter current, and the collector current is actually equal to the sum of the base and emitter currents. However, as the base current is very low in comparison to the emitter current there is only a negligible difference between the collector and emitter currents.

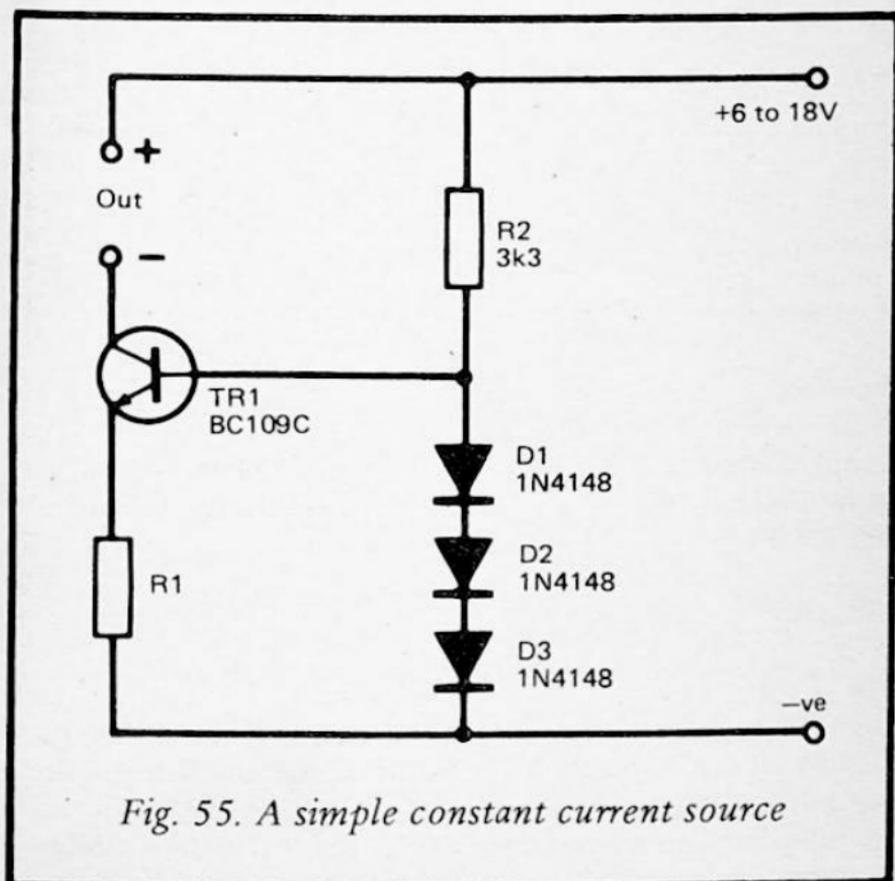


Fig. 55. A simple constant current source

Thus, provided the load connected in the collector circuit of Tr1 has a sufficiently low impedance (since there is a limit to the supply voltage available to force a current through the load) the current flow will be virtually the same whatever that impedance happens to be. The value of R1 is selected to give the desired current flow, and the required value is calculated using the following formula:-

$$R1 = 1.4/I \text{ out ohms}$$

For example, a current of 2mA (0.002 amps) would require R1 to have a value of 1.4 divided by 0.002, which equals 700 ohms. Of course, this can only be approximate, and where it is necessary to have a very high degree of accuracy R1 should be a preset resistor so that the output current can be adjusted to precisely the required level.

This circuit is suitable for output currents from a few tens of microamps to about 20mA or so. It cannot be used at currents much higher than 20mA as Tr1 might then overheat. For higher currents Tr1 should be replaced with a Darlington power transistor such as an TIP122, and an additional diode should be added in series with D1 to D3 (to compensate for the higher base - emitter voltage drop of a Darlington device). The circuit will then provide currents of up to a few hundred milliamps provided Tr1 is given adequate heatsinking.

Figure 56 shows an alternative constant current generator where one side of the load connects to the negative supply rather than the positive one. This circuit is essentially the same in operation as the previous one, but uses a PNP transistor instead of an NPN one. For a high current version the TIP127 is a suitable device to use in the Tr1 position.

Simple Latch

The circuit of Figure 57 has an output that is normally low (at about +2 volts) but which goes high (about 1 volt less than the positive supply voltage) if the input is taken more than about 0.6 volts above half the supply potential. Once the output has triggered to the high state it will remain in that state until the reset switch is operated or the supply is switched off. In other

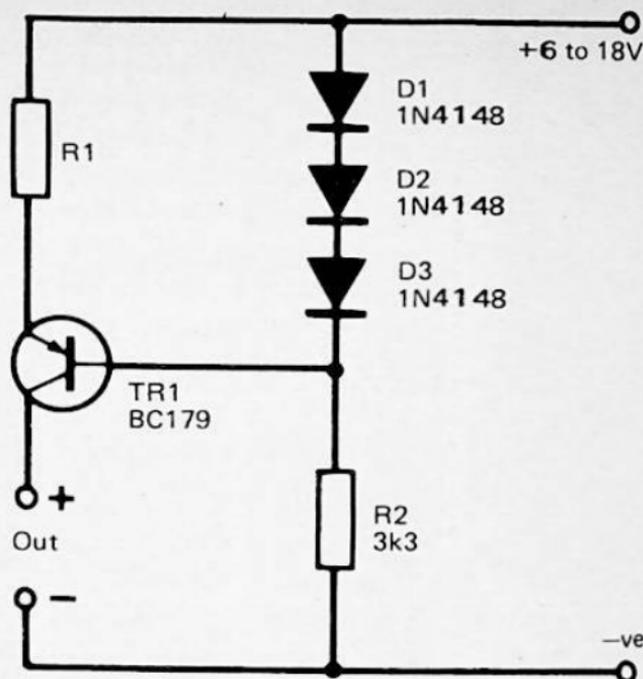


Fig. 56. An alternative constant current generator circuit

words a brief input signal will give an output pulse of indefinite length.

An operational amplifier used as a comparator is the basis of the circuit, and the inverting input of IC1 is biased by R1 and R2 to half the supply voltage. However, at switch-on C1 provides a positive pulse to the inverting input. This ensures that the output initially assumes the low state, and due to the positive feedback through R3 it latches in this state.

If the input of the circuit is taken more than about 0.6 volts above the bias voltage at the inverting input the non-inverting input will be the one at the higher potential and the output of IC1 will trigger to the high state. R3 provides positive feedback which ensures that the output triggers rapidly to the high state and latches in that state. Even if the input is now taken well

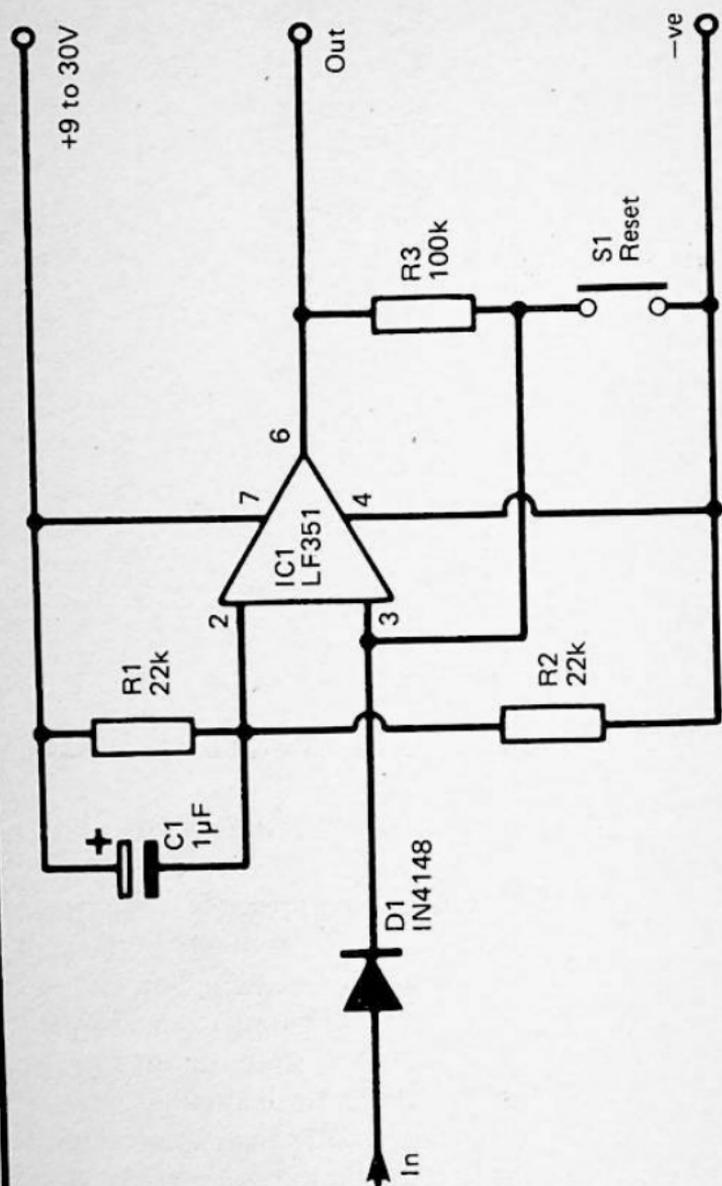


Fig. 57. A simple latch circuit

below half the supply voltage the output will stay in the high state since D1 will prevent the input signal from pulling the non-inverting input of IC1 lower in voltage. D1 only allows the input to increase the voltage at the non-inverting input.

The circuit can be reset to the low output state by briefly operating S1 so that the non-inverting input is taken to the negative supply potential, the output goes low, and the positive feedback through R3 ensures it remains in that state when S1 is released.

The current consumption of the circuit is about 2 to 3mA.

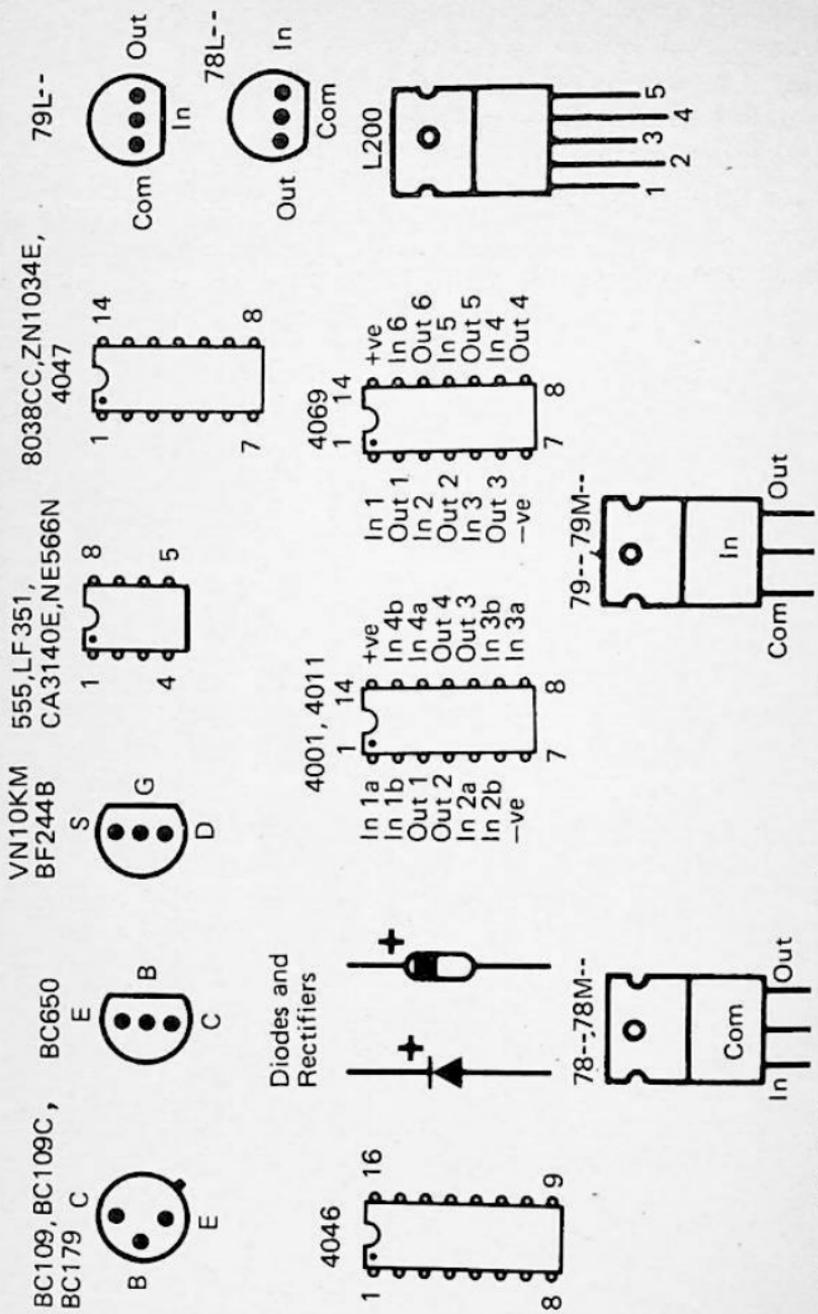


Fig. 58. I.C. top views and transistor base views

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