

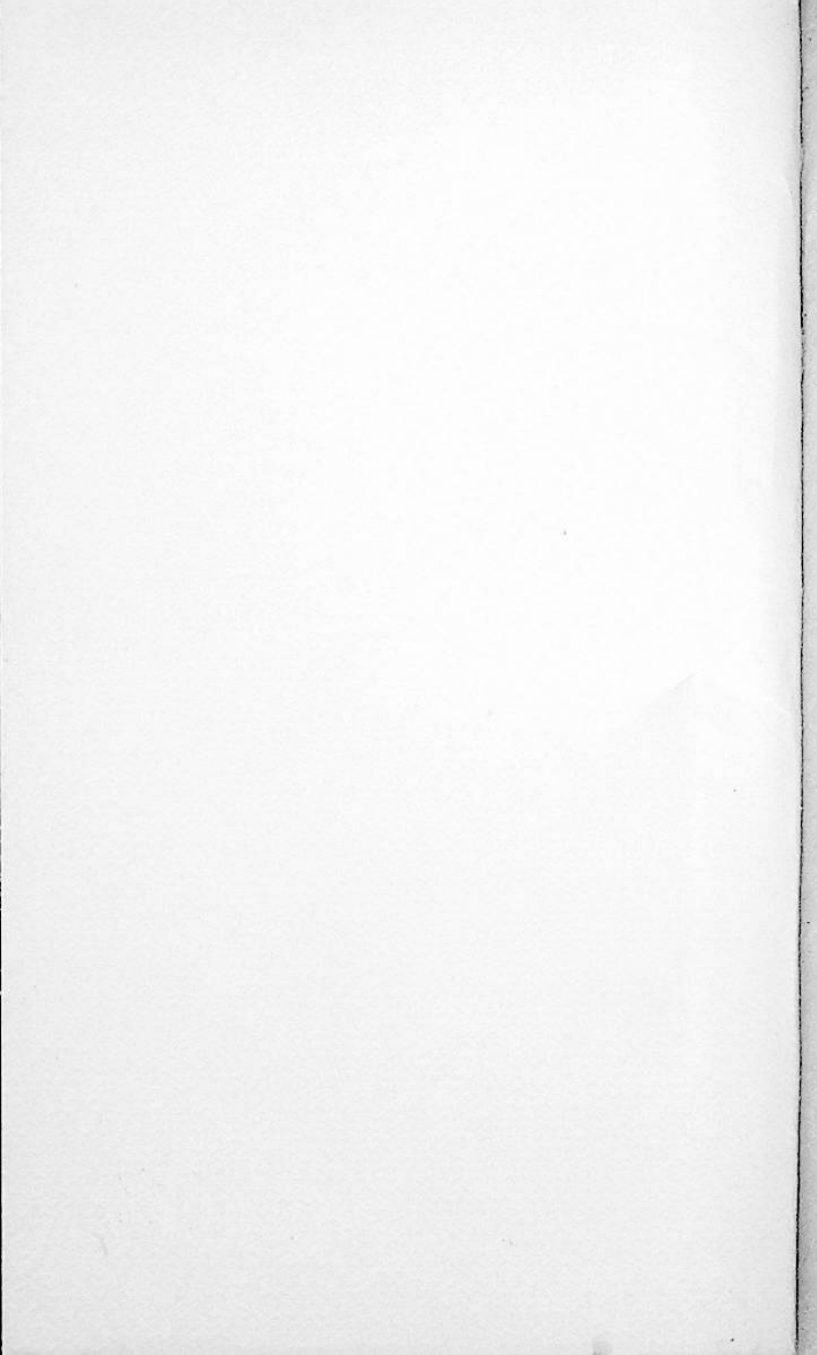
VMOS Projects

R. A. PENFOLD



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VMOS PROJECTS

by
R. A. PENFOLD

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CHAPTER 1

INTRODUCTION

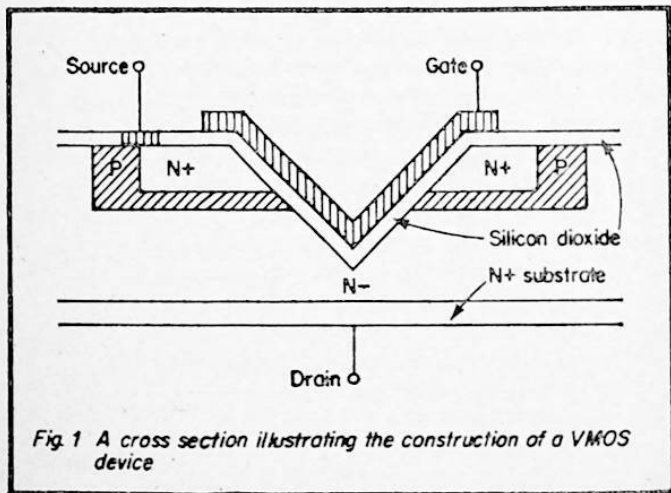
Although modern bipolar power transistors are far superior to the earlier germanium devices and are capable of giving excellent results in a wide range of applications, they are not without their drawbacks. The main problems are secondary breakdown, thermal runaway, limited high frequency response and a fairly high driving power requirement.

With the advent of field effect devices it seemed that it would only be a matter of time before improved power devices without these disadvantages became available. This has in fact proved to be the case, although it has taken a number of years for the power FET to be produced commercially. However, a number of different devices are now available from component retailers and it seems likely the range of devices will increase as improved types are developed, and the advantages of power FETs cause them to increase in popularity.

There are three basic types of power FETs; the VMOS type, power MOSFETs, and V-JFETs. This book will primarily be concerned with VMOS power FETs because these are readily available at reasonably low prices. Low, medium and high power devices can be obtained, and although only N channel devices were originally produced, P channel devices are now produced commercially as well. Power MOSFETs are available to the amateur user, although they are relatively expensive. These are primarily designed for use in audio amplifiers (but are also suitable for a variety of other applications), and will be dealt with in the chapter dealing with audio circuits. Like power MOSFETs, V-JFETs are mainly intended for use in high quality high power audio amplifiers. However, they are depletion mode devices, whereas VMOS and power MOSFETs are enhancement mode devices. In other words, a V-JFET is switched hard on when it has a gate to source voltage of zero, and is switched off by reverse biasing its gate (as in the case of an ordinary JFET). VMOS and power MOSFETs are switched off if the gate to source voltage is made zero, and are switched on by applying a forward bias to the gate. In this respect they are

like ordinary bipolar transistors, or the MOSFETs fabricated in CMOS ICs. V-JFETs are more difficult to use than the other two types of power FET, and would not seem to be available to amateur users. Therefore, they will not be considered further here.

An ordinary field effect transistor has an "on" resistance that is usually in the range 100 to 500 ohms, and so producing a useable power FET is not just a matter of producing a device that is capable of dissipating high power without being destroyed it is also necessary to obtain a much lower "on" resistance. The structure of a VMOS FET is therefore substantially different to that of an ordinary MOSFET, and Figure 1 shows a cross-section that illustrates the internal structure of a VMOS device. With an ordinary MOSFET (or JFET) the current flows horizontally between the drain and source terminals, but the structure of a VMOS device is such that this current flow is vertical, and a low "on" resistance is obtained. Another necessary characteristic of VMOS devices is that they have a far higher gain than conventional FETs, so that a gate signal of a few volts peak to peak is sufficient to produce a change in drain current of perhaps an ampere or more.



Advantages

The most obvious advantage of a VMOS device over an ordinary bipolar power transistor is that a VMOS device requires no significant input current, and has a typical input impedance of thousands of megohms. The input current is of the order of a few nano-amps, rather than milli-amps, and a high impedance drive circuit is quite adequate except where fast switching speeds are required. VMOS devices have an input capacitance of up to about 50pF, and in high speed switching or similar applications the drive circuit must be at a low enough impedance for this to have no significant effect.

VMOS devices are ideal for high speed applications since they do not have the minority carrier storage time associated with bipolar transistors, and are majority carrier devices. The typical rise and fall time for a VMOS transistor is only about 4 nano-seconds, which is about 10 to 100 times less than a bipolar power transistor.

An important advantage of VMOS devices in some applications, such as class B audio power amplifiers, is that they have a negative temperature coefficient and do not suffer from thermal runaway. Bipolar transistors have a positive temperature coefficient, and thus when they heat up in use they tend to conduct more heavily. This can produce increased current flow and dissipation in the device, which causes further heating and a further increase in current flow. This process can continue until the device becomes overheated and is destroyed. Since VMOS transistors have a negative temperature coefficient, when they heat up they tend to conduct less heavily, giving slightly reduced current flow and no possibility of thermal runaway. This is not to say that a VMOS device cannot be destroyed by overheating, since the reduction in current flow due to heating is not very large, and is not sufficient to make it impossible to over-power the device.

A further important advantage of VMOS devices is that they do not suffer from secondary breakdown. In bipolar transistors this effect is caused by a sort of local thermal runaway causing "hot spots" which cause the silicon to melt and produce a short circuit between the collector and emitter

terminals! To avoid this problem it is necessary to operate the device at safe combinations of collector current and voltage, and the manufacturers data shows the "area of safe operation". The structure and negative temperature coefficient of VMOS devices eliminates secondary breakdown, and VMOS transistors can operate with virtually any combination of drain voltage and current (provided the maximum permissible voltage, current, and power ratings are not exceeded, of course).

Disadvantages

VMOS devices are not perfect, and do have a couple of disadvantages, although these are of a relatively minor nature. One is simply that they are slightly more expensive than bipolar power devices, although this is offset by the fact that in many application it would take two or even three bipolar devices (plus possibly a few passive components as well) to replace a VMOS transistor, and is perhaps not a valid criticism. The second disadvantage is that most of the current devices have a comparatively high "on" resistance. This means that the voltage developed across a VMOS device when switched hard on and passing a high current is likely to be somewhat more than would be the case for a bipolar device operating under similar conditions. This can result in reduced efficiency in some applications. However, devices having "on" resistances of only about 0.2 or 0.3 ohms are now produced, and two or more devices can be connected in parallel, as shown in Figure 2, to give reduced "on" resistance. Although this simple parallel method of connection does not work well with bipolar transistors, where the positive temperature coefficient can result in one device passing a steadily increasing percentage of the total current flow, until it eventually overheats and is destroyed, this does not occur with VMOS transistors due to their negative temperature coefficient. In fact the circuit tends to be self stabilising with the two devices taking a virtually identical share of the current flow. Of course, the transistors used should all be of the same type, or should have similar characteristics, or a greatly unequal current could flow through them, and the stabilising action might be inadequate to properly compensate for this. The "on" resistance is equal to

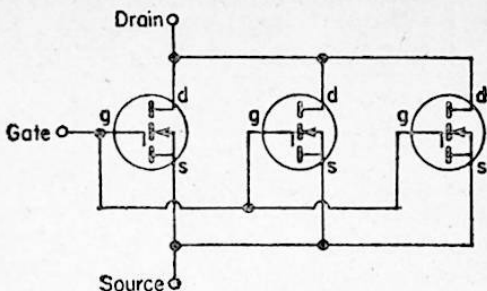


Fig. 2 VMOS devices can be added in parallel to reduce "on" resistance

the "on" resistance of a single device divided by the number of devices used,

Available Types

The table given below lists most of the types available to amateur users at the time of writing, and gives the principle characteristics of these devices.

Device	Maximum Power Dissipated	Maximum Drain Current	Maximum Drain-Source Voltage	Typical "on" Resistance
VN46AF	15W	2A	40V	2 ohms
VN66AF	15W	2A	60V	2 ohms
VN67AF	15W	2A	60V	2 ohms
VN88AF	15W	2A	80V	3 ohms
VN10KM	1W	0.5A	60V	4 ohms
VN64GA	80W	12.5A	60V	.0.3 ohms
VMP4	25W	2A	60V	2 ohms

The above are all N channel devices, but a complementary pair are available, and these have the parameters shown in the table provided below.

<i>Device</i>	<i>Maximum Power Dissipated</i>	<i>Maximum Drain Current</i>	<i>Maximum Drain-Source Voltage</i>	<i>Typical "on" Resistance</i>
BD512	10W	1.5A	60	3.5 ohms
BD522	10W	1.5A	60	2.5 ohms

The BD512 is the P channel device, and the BD522 is the N channel one. These devices do not have any protection against static discharge, and neither do the VMP4 and VN64GA from the previous table. The other devices have a 15 volt zener diode connected across the source and gate terminals to limit the maximum forward gate voltage to 15 volts, and give protection against static charges. It is unlikely that one of the protected devices would be damaged by static discharge, but the unprotected devices must be handled with some care. It is advisable to make these the last components to be connected into circuit, and to short circuit their terminals (say with a piece of aluminium foil) until they have been wired into place.

Care must be taken when using the protected devices, as a high gate current will flow (due to the internal zener) if the forward gate to source voltage should even slightly exceed 15 volts, and this could easily lead to the destruction of the device. The same thing can happen if the gate is slightly reverse biased (the zener then acting rather like an ordinary forward biased diode). With unprotected devices a maximum gate to source voltage of ± 30 volts is permissible.

Output Characteristic

Although it is normal to refer to the "on resistance" of a VMOS transistor, it should perhaps be pointed out that it is not a true resistance that is provided between the drain and source terminals of the device. Up to a point, there is a virtually linear relationship between applied voltage and the current flowing through the device, and the device does provide what for all practical purposes can be regarded as true resistance. However, above a certain threshold level, increased drain to source voltage has little effect on the current passed by the device. The greater the gate bias supplied to the

transistor, the greater the threshold current. This effect is shown in Figure 3 which shows the output characteristic of a typical VMOS device. This is similar to the output characteristic of normal FETs, but the currents involved are, of course, very much higher.

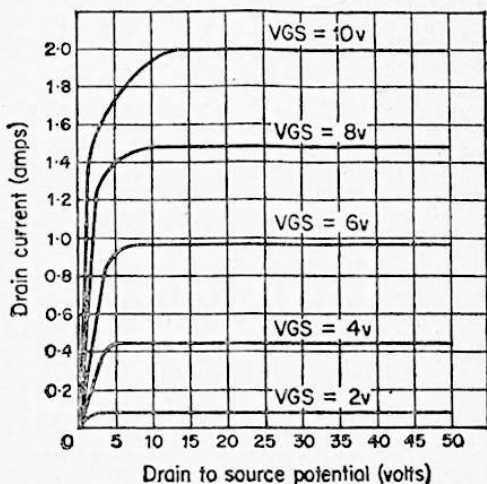


Fig. 3 Typical output characteristic of a VMOS transistor

When designing circuits using VMOS transistors it should be borne in mind that the gate voltage versus drain current characteristic of a VMOS device is very different to the base voltage versus collector current characteristic of a bipolar transistor. Simply changing a bipolar transistor (or Darlington Pair) for a VMOS device may not necessarily give good results, and could in certain circumstances fail to work at all. A bipolar transistor does not start to conduct significantly until the base to emitter voltage equals about 0.5 volts or so (for a normal silicon device anyway), and an increase of only about 200mV or so is needed in order to bias the device into saturation,

The threshold voltage of a VMOS device is somewhat more variable, but is higher, usually being between about 0.8 and 2 volts. The gate potential needed to bias the device into saturation is very much higher than the threshold voltage, and would typically be something in the region of 10 volts. This can be seen by referring to Figure 4 which shows the transfer characteristic for a typical VMOS transistor.

Although in terms of input voltage to output current a VMOS device may seem to be considerably inferior to a bipolar device, and is in fact inferior, it should of course be borne in mind that at saturation a bipolar device may well require an input current to 100mA in a high power circuit. A VMOS device requires no significant current at all!

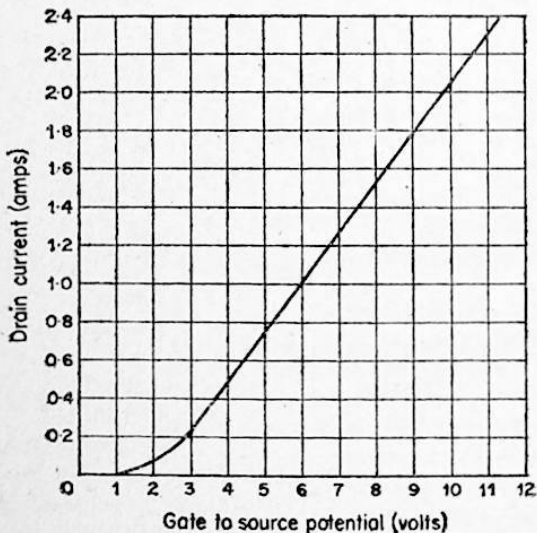


Fig. 4 Typical transfer characteristic of a VMOS transistor

CHAPTER 2

AUDIO CIRCUITS

A number of audio power amplifier circuits using either VMOS or power MOSFETs will be described in this chapter. Preamplifier circuits and other low level audio circuits will not be considered, as although VMOS devices could be used in such applications, they would not perform as well as bipolar and JFET devices, and are not really intended for such uses. They are designed for good results in medium and high power applications, rather than for low noise and distortion in low power audio circuits.

200mW Class B Amplifier

Probably the lowest power stage for which a VMOS device is a practical proposition is as the driver stage for a small audio power amplifier. The circuit of Figure 5 shows the circuit of such an amplifier.

This uses what is basically a conventional configuration having a Darlington pair emitter follower driver stage, and a complementary emitter follower output stage. However, the Darlington pair driver stage actually uses a VMOS device instead. This is Tr1, and a VN10KM device is more than adequate here since the maximum power and current handled by this stage are both quite low (only about 18mW and 8mA respectively). Larger devices such as the VN67AF will work in the circuit though.

R1 and R2 are used to bias the output of the amplifier to about half the supply voltage. This can only be approximate though, since the gate voltage needed to give the appropriate quiescent drain current of 4mA in Tr1 will vary somewhat from one device to another. For optimum results R2 should be changed for a 2.2 megohm preset resistor so that the quiescent output voltage can be adjusted to half the supply potential.

R3 is the drain load for Tr1, and D1 plus D2 are included in the drain circuit of Tr1 to provide a quiescent bias to the

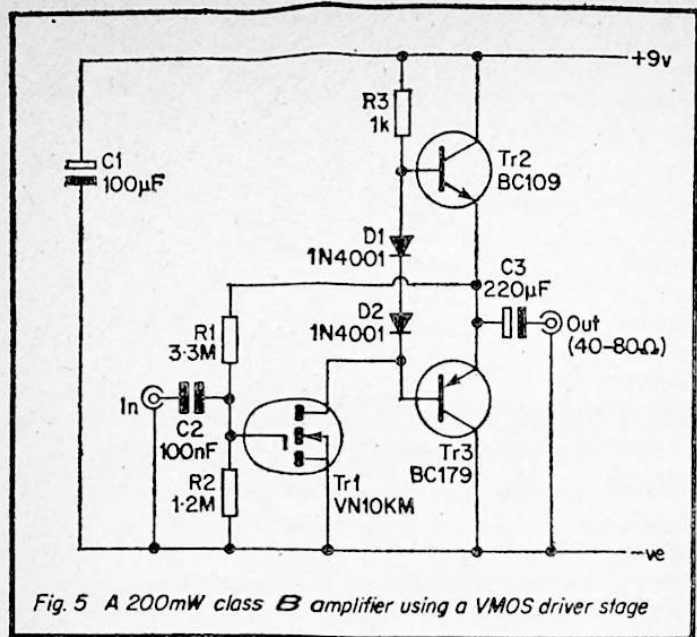


Fig. 5 A 200mW class B amplifier using a VMOS driver stage

output transistors, Tr2 and Tr3. This quiescent bias is needed because about 0.5 volts or so is needed across the base and emitter terminals of a silicon bipolar transistor before it starts to conduct between its collector and emitter terminals. If the bases of Tr2 and Tr3 were simply connected together and fed from the drain terminal of Tr1, a voltage swing of as much as 1 volt peak to peak (± 0.5 volts) would not cause either of the output transistors to be biased into conduction, and would not produce any significant change in the output voltage! Larger signal levels would bias the output devices into conduction for a proportion of the time, and the output would swing either side of its quiescent level during signal peaks. However, the signal at the output would be severely distorted to say the least, with the beginning and end of each half cycle being totally absent.

Ideally, the output stage should be biased so that both the output devices are passing a reasonably large quiescent current.

A low quiescent current has the disadvantage of higher distortion because the output transistors do not respond to even small voltage swings from the driver stage, but not as well as at higher signal levels. This is merely because a bipolar transistor tends to have a relatively low current gain at low levels of collector current. Thus, even with a small quiescent bias current, the distortion (which is termed "crossover distortion") is still produced, although much less severely than with zero quiescent bias.

In this design D1 and D2 produce a bias voltage that is just sufficient to bring Tr2 and Tr3 to the threshold of conduction, so that a small quiescent output current is produced through these devices. A higher current would bring about two main problems. Firstly, a small amplifier such as this is likely to be used mainly in applications where it will be powered from an ordinary 9 volt dry battery, and in the interests of battery economy a fairly low quiescent current is desirable. The second problem is that of thermal runaway in the output transistors. It has already been pointed out that bipolar transistors have a positive temperature coefficient, and therefore tend to conduct more heavily as they heat up. This heating will inevitably occur in the output devices in even a small amplifier such as this, and can lead to a substantial increase in the quiescent current consumption after the unit has been in use for a while. The increased bias current then leads to further heating of the output transistors, which in turn results in further heating, and so on. This can easily result in a steady increase in the output current until the dissipation in the output devices becomes so great that they overheat completely and are consequently destroyed.

This problem can be alleviated to some extent by using a very low quiescent bias current through the output devices, so that even after the output devices have heated up slightly with use, there is still only a very modest quiescent current flow, and no danger of thermal runaway. The price that has to be paid for this simple method is an increase in crossover distortion. However, the use of negative feedback reduces this distortion to a suitably low level, and a small amplifier such as this is not really intended to give hi-fi results anyway.

One other problem with the stability of the quiescent

bias current is that it would tend to fall somewhat with low ambient temperatures, and increase at high ambient temperatures. This could cause problems with either excessive crossover distortion or reduced thermal stability. This is overcome here (and in most other designs) by the use of silicon diodes to provide the bias voltage for the output transistors. These give a bias voltage that increases at low ambient temperatures, and decreases at high ambient temperatures. This tends to stabilise the bias current against changes due to variations in ambient temperature.

In more powerful amplifiers where very high quality is required, and a fairly substantial quiescent output current must be used in order to minimise crossover distortion, the bias diodes (or a transistor connected as an "amplified diode") are often mounted close to, or actually on the heatsink for the output transistors. The diodes then sense the rise in temperature of the output devices, and reduce the bias voltage to them so that thermal runaway is avoided. As we shall see shortly, problems of this type do not occur with VMOS and power MOSFET transistors, allowing the use of much more simple but nevertheless more reliable biasing methods.

This simple circuit has the usual emitter follower output stage with Tr2 being used to drive the speaker during positive going output excursions, and Tr3 driving it during negative output excursions. C3 provides DC blocking at the output, which is suitable for a speaker having an impedance in the range 40 to 80 ohms. The use of a lower speaker impedance is not recommended, as this would result in increased heating in the output transistors, which could in turn lead to problems with thermal stability. The circuit is a Class B type, and although the quiescent current consumption is only about 5 or 6mA, the current consumption increases to about 20 or 30mA at high volume settings. The output power of the circuit is about 100mW RMS or so into an 80 ohm speaker, or around 200mW RMS into a 40 ohm load. Approximately 500mV RMS is needed at the input to produce maximum output. The input impedance is quite high at about 1 megohm.

Components: 200mW Class B Amplifier (Figure 5)

Resistors (all 1/3 watt 5% (10% over 1M))

R1	3.3M	R3	1k
R2	1.2M		

Capacitors

C1	100 μ F 10V	C3	220 μ F 10V
C2	100nF plastic foil		

Semiconductors

Tr1	VN10KN (or VN66AF, VN67AF, etc.)	Tr3	BC179
Tr2	BC109		
D1	1N4001	D2	1N4001

Miscellaneous

Component panel, high impedance speaker, wire, solder, etc.

Complementary VMOS Amplifier

Now that P channel VMOS devices are available in addition to the original N channel devices, it is possible to produce a true complementary Class B VMOS design using circuitry similar to that employed in bipolar circuits of this type. However, the characteristics of VMOS devices enable some simplifications without any loss of performance or reliability.

The circuit diagram of a simple true complementary VMOS amplifier is shown in Figure 6. This uses a well known arrangement having a common emitter input stage (Tr1) directly driving a common emitter driver stage (Tr2). This in turn directly drives a pair of complementary emitter follower output transistors, or complementary source follower output devices as they are in this case, of course. There is virtually 100% negative feedback at DC due to the feedback provided from the output of the amplifier (Tr3 and Tr4 sources) to the emitter of Tr1 by R6. This makes it easy to bias the output to nominally half the supply potential, since it is merely necessary to bias the input (Tr1's base) to this level using a potential divider. The bias circuit consists of R1, R2, R3 and C2. R1 and C2 are included to filter out any hum or noise that might otherwise be coupled from the supply lines to the input of the amplifier via the bias circuitry. Normally the output would be biased to half the supply potential in order to give the

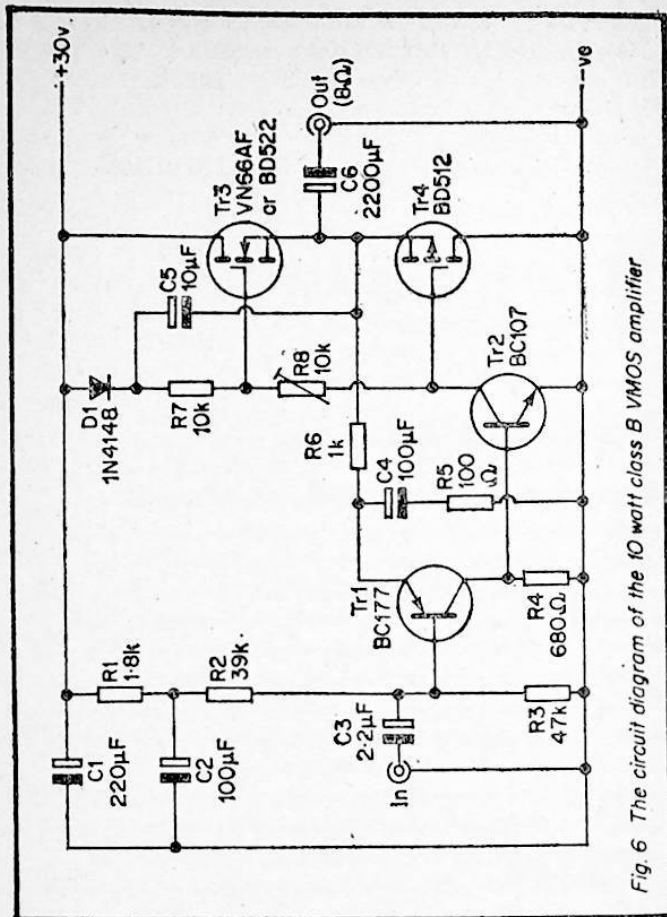


Fig. 6 The circuit diagram of the 10 watt class B VMOS amplifier

highest possible output voltage swing (and hence power also) before the onset of clipping and serious distortion. As we shall see shortly, optimum results in this case are obtained with a slightly higher quiescent output voltage, and so the unit is biased accordingly.

R4 is the collector load for Tr1, and has a value which sets the collector current of Tr1 at a suitable level of about 1mA. R7 is the main collector load for Tr2, and R8 also forms part

of Tr2's collector load. The latter sets the quiescent bias current through the output transistors, and is adjusted for a total quiescent current consumption of about 30mA. It is important that this component should be adjusted for virtually minimum value when the amplifier is first connected to the power source, as otherwise a very high initial current might be drawn by the unit. This could possible damage Tr3 and (or) Tr4, and could damage the power supply as well. With R8 set for virtually minimum resistance there should be no significant current through the output devices, and R8 can then be carefully advanced to give the appropriate current consumption figure.

Although in a design which uses bipolar devices it would be normal to use a transistor or diodes in the bias circuit for the output transistors (as explained in the previous section of this book) to provide temperature compensation, this is unnecessary here of course, since the output devices both have a negative temperature coefficient, and will cause a slight reduction in the quiescent outout current when they heat up in use. There is thus no risk of thermal runaway, and no need for temperature compensation since the small drop in the output bias current when the output transistors heat up is of no practical consequence.

One disadvantage of VMOS devices used in this type of circuit is a lower efficiency when compared to bipolar devices in an equivalent circuit. When used in the emiiter follower mode, the input voltage at the base of the transistor produces an almost identical change in the output voltage at the emitter, and there is a voltage drop of only about 0.65 volts between the base and emitter terminals of the device. The voltage gain is not quite unity, but is normally about 0.98 and quite good enough to give high efficiency in a circuit of the type under consideration here, with an output voltage swing that is not far short of being equal to the supply potential if the amplifier is fully driven.

A VMOS device has a gate threshold voltage that is somewhat higher than the equivalent voltage in a bipolar device, and this gives an output voltage swing that is accordingly lower. However, this is largely offset by the fact that in bipolar power amplifiers it is often necessary to use a Darlington

pair or some similar arrangement for each output device, and this often effectively doubles the base threshold voltage. VMOS devices are still less efficient though, since they need a substantial gate to source voltage to bias them hard into conduction, whereas a bipolar device needs a base to emitter voltage that is only fractionally higher than its base threshold voltage to achieve a similar state of conduction. Thus a VMOS device provides substantially less than unity voltage gain when used in the source follower mode, and the voltage drop from the gate to the source becomes increasingly large as the output current increases.

One way of minimising this problem is to use the bootstrapping technique, and in this design the bootstrapping is provided by C5 and D1. Under quiescent conditions D1 enables current to pass through R7, R8 and Tr2, and it therefore has little effect on the circuit. When the output is positive going, C5 couples this increase in voltage to the junction of D1 and R7. Thus, on positive going output signals the supply voltage to the driver stage is effectively increased by an amount equal to the change in output potential (minus the voltage drop of about 0.5 volts across D1). D1 must be included so that the bootstrapping signal is isolated from the positive supply rail. It is quite common to find a resistor in the D1 position in this type of circuit (which is also used extensively in bipolar designs to optimise the output power), but if a resistor were to be used C5 would be looking into a lower impedance and would therefore need to be higher in value to ensure good results at low frequencies.

Of course, the point of using the bootstrapping technique is that it enables Tr3 to have a gate voltage which is in excess of the positive supply potential, and even with a voltage drop of several volts from its gate to its source terminals it can have a source potential that is very nearly equal to the positive rail voltage. In this case the peak positive output voltage will in fact still be substantially short of the positive supply voltage since the output feeds into a nominal 8 ohm load, and the typical "on" resistance of Tr3 is either 2 ohms (VN66AF) or 2.5 ohms (BD522). By a potential divider action this obviously gives a significant voltage drop across the output transistors.

One drawback of the bootstrapping technique is simply

that it only gives the desired effect on one set of half cycles; the positive going half cycles in this case. The minimum drive voltage available to Tr4 remains 0V, or a little higher than this in reality since the collector voltage of Tr2 cannot quite go right down to the negative supply voltage. With a minimum gate voltage of just over 0V, the minimum source voltage of Tr4 must be several volts lower than this under high output current conditions.

In order to compensate for this it is necessary to use a slightly higher supply voltage than would normally be utilized for an amplifier of this power rating and load impedance. Also, as mentioned earlier, the output is biased to slightly more than half the supply potential, this being necessary because the output would otherwise clip on negative peaks well before the point where clipping of positive output peaks occurred. Biasing the output a few volts higher than half the supply voltage ensures virtually symmetrical clipping, and optimum output power for a given supply voltage.

Although a voltage gain of unity is very convenient when it comes to correctly biasing the amplifier, such a low gain is, of course, totally inadequate for use in practical audio applications. It is therefore necessary to decouple some of the feedback at audio frequencies so as to give a useable level of voltage gain. C4 is used to provide this decoupling, and R5 is included in series with C4 in order to limit the amount of negative feedback that is removed. This is done because the full voltage gain of the amplifier is not required in the majority of applications, and it is beneficial to leave as much negative feedback as possible as this feedback reduces distortion.

The voltage gain of the amplifier is approximately equal to R6 divided by R5, or about 10 times (20dB) with the specified values. However, within reason the voltage gain of the circuit can be altered to suit individual requirements by modifying the value of R5.

As with any audio amplifier design, the performance will vary somewhat from one example of the design to another. However, the distortion performance of this design seems to be comparable to that of simple bipolar transistor circuits, with the total harmonic distortion at most power levels being only about 0.1% or less. Measurements were only made at a

frequency of 1kHz, but the open loop gain of the circuit (i.e. the voltage without any negative feedback applied) is virtually constant over the entire audio frequency band, and so little fall off at the higher audio frequencies would be expected. Those readers who are familiar with bipolar designs of this type may be surprised at the lack of any components in this design to provide high frequency roll-off or phase compensation. Circuitry of this type is normally necessary in bipolar circuits due to the comparatively poor high frequency response of the output transistors giving rise to significant phase lags at high frequencies. This can result in the feedback through the negative feedback loop actually being positive rather than negative at high frequencies, and if the gain of the circuit is sufficient at these frequencies the inevitable result is oscillation. The normal way of preventing this is to use phase compensation networks and (or) high frequency roll off in the circuit. This results in an open loop gain which reduces at the higher audio frequencies, giving reduced feedback and performance at these frequencies.

Similar problems do not usually arise in VMOS audio circuits because the frequency response of VMOS devices is extremely good, and the phase lags that cause the instability are simply not generated. This gives a level of performance at high frequencies which is not significantly different from that at low and middle frequencies.

As is the case with most amplifiers, the level of distortion rises significantly as the onset of clipping approaches, and rises very steeply once clipping actually commences.

The output power of the circuit is typically about 6 to 7 watts RMS into an 8 ohm load when using a 30 volt (loaded) supply. A higher output power of up to about 10 watts RMS can be obtained using a supply of about 36 volts, and this is about the maximum loaded supply voltage that should be used.

If pairs of parallel connected devices are used in the output stage it is possible to obtain a somewhat higher output power for a given supply voltage. About 8 to 9 watts RMS is available using a 30 volt supply, and up to about 13 watts RMS using a 36 volt supply.

Note that a considerable amount of heat is produced in the output devices (especially Tr4) and they must be mounted on

substantial heatsinks.

Components: 10 Watt Class B VMOS Amplifier (Figure 6)

Resistors (all 1/3 watt 5% (10% over 1M), except R8))

R1	1.8k	R5	100 ohms
R2	39k	R6	1k
R3	47k	R7	10k
R4	680 ohms	R8	10k 0.25W preset

Capacitors

C1	220 μ F 40V	C4	100 μ F 25V
C2	100 μ F 40V	C5	10 μ 25V
C3	2.2 μ F 25V	C6	2200 μ F 40V

Semiconductors

Tr1	BC177	Tr3	BD522 or VN66AF
Tr2	BC107	Tr4	BD512
D1	1N4148		

Miscellaneous

Component board, 8 ohm speaker having a rating of at least 10 watts RMS, wire, solder, etc.

Quasi-Complementary VMOS

Early bipolar class B amplifier designs used quasi-complementary arrangements in the output stage, and it is possible to obtain good results using VMOS devices in similar arrangements.

Figure 7 shows the circuit diagram of a simple quasi-complementary VMOS power amplifier using three N channel VMOS devices plus an operational amplifier input stage.

The output stage uses Tr1 to Tr3 in an arrangement that is commonly known as a "Totem Pole" output stage. This is often employed in the output stages of logic devices, but is not much used in audio amplifiers due to the production of large amounts of crossover distortion, as we shall see shortly.

If we consider only the output section of the circuit for the moment (Tr1 and the components to its right), with the input to this circuitry (Tr1 and Tr3 gates) at the negative supply potential, Tr1 and Tr3 will both be switched off. Tr2 on the other hand, has its gate terminal fed from the drain of Tr1, and this will be at virtually the full supply potential with Tr1 switched off. Tr2 is therefore biased hard into conduction

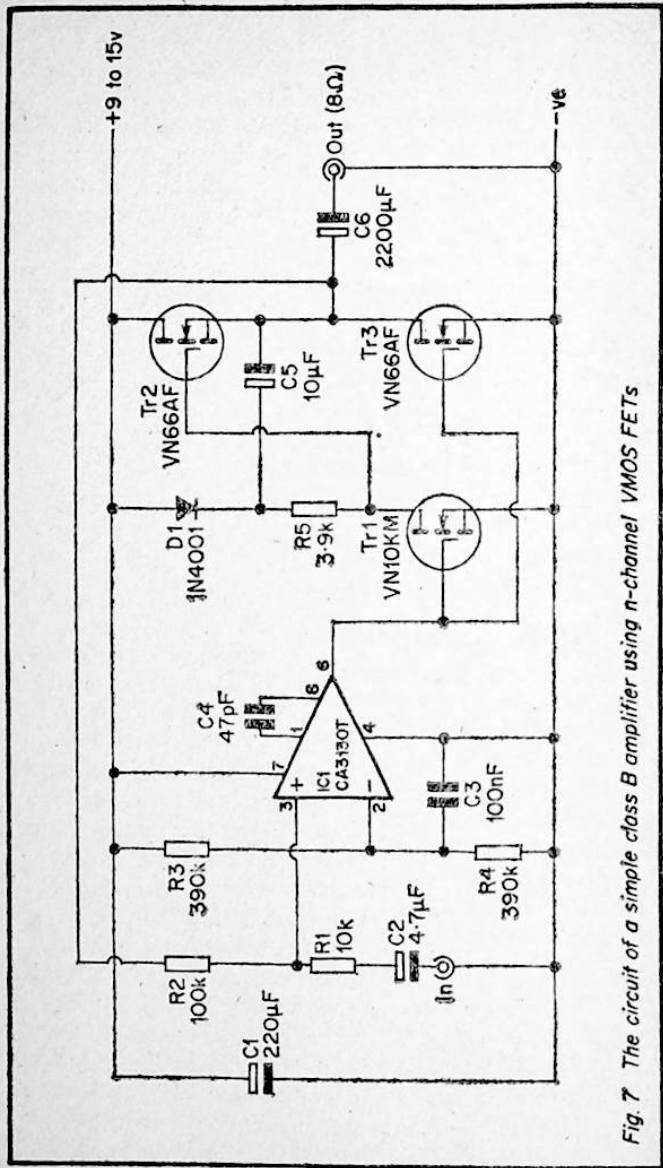


Fig. 7 The circuit of a simple class B amplifier using n-channel VMOS FETs

and the output goes to virtually the full positive supply potential.

If the input voltage is gradually increased, at some point Tr1 and Tr3 will start to conduct. As Tr1 has a fairly high load impedance in its drain circuit it exhibits a substantial voltage gain, and its drain voltage rapidly falls. This causes Tr2 to switch off and the output to swing negative. Further increasing the input voltage causes Tr3 to conduct heavily and pull the output voltage down to virtually the negative supply potential. Reducing the input voltage back to zero causes this whole process to be repeated in reverse, so that the output voltage swings back to almost the full positive supply voltage. Thus, by varying the input voltage the required push-pull action is obtained in output transistors Tr2 and Tr3.

The problem with this system is that as the input voltage increases and Tr2 switches off, Tr3 only starts to switch on relatively slowly. This ensures that the circuit does not go through a period when both output transistors are switched on fairly hard simultaneously causing a large current to flow through them. It does have the disadvantage though, of giving an effect similar to conventional cross over distortion. There is also the problem of an effective mismatch in the output transistors as the gain of Tr3 in the lower section of the output stage is likely to be less than the combined gain of Tr1 and Tr2 in the upper part of the output stage.

This makes the "Totem Pole" arrangement unsuitable for use in applications where the highest possible quality is required (in this basic arrangement anyway), but it is suitable for low power applications where something less than true hi-fi reproduction is perfectly acceptable.

The relatively high distortion in the output stage must be counteracted by the use of fairly substantial amounts of negative feedback. The output circuitry is driven by a high gain operational amplifier with overall negative feedback from the output of the amplifier to the non-inverting (+) input by way of R2. Although negative feedback would normally be taken to the inverting input of an operational amplifier, it must be applied to the non-inverting input in this case due to the phase inversion through the output circuitry. R1 and R2 set the voltage gain of the circuit at nominally ten times (20dB)

and this ensures a large amount of negative feedback is applied to the circuit since the gain of IC1 is considerably higher than this at audio frequencies. R1 also sets the input impedance of the amplifier; the actual figure being about 10k.

R3 and R4 bias the inverting input (and also the output of the amplifier) to about half the supply potential. C4 is the compensation capacitor for IC1, and aids the stability of this device. D1 and C5 provide bootstrapping which helps to give good efficiency from the circuit. Since the lower transistor in the output stage (Tr3) is used in the common source mode and not the source follower mode, the circuit also operates efficiently on negative output peaks. This is essential in a circuit which operates from a relatively low supply voltage as poor efficiency would also result in only a very limited output voltage swing and output power.

The output power of the circuit is about 800mW RMS using a 9 volt supply and an 8 ohm loudspeaker, or a little over 2 watts RMS using a 15 volt supply and an 8 ohm load. 15 volts is the maximum supply voltage that should be used. The quiescent current consumption is only a few mA, but a supply current of about 250mA is drawn with an output power of 2 watts RMS.

Although the dissipation in the output transistors is not very high it would be advisable to fit them with small finned heatsinks. This will give them some protection against overheating in the event of a short circuit or overload. It should also be remembered that the "on" resistance of a VMOS device rises somewhat with increased temperature, and keeping the output devices as cool as possible ensures optimum efficiency.

Components: Simple Class B VMOS Amplifier (Figure 7)

Resistors (all 1/3 watt 5%)

R1	10k	R4	390k
R2	100k	R5	3.9k
R3	390k		

Capacitors

C1	220 μ F 25V	C4	47pF ceramic plate
C2	4.7 μ F 25V	C5	10 μ F 25V
C3	100nF plastic foil	C6	2200 μ F 25V

Semiconductors

IC1 CA3130T (or CA3130E)
Tr1 VN10KM (or VN66AF, Tr3 VN66AF
VN67AF, etc)
Tr2 VN66AF
D1 1N4001

Miscellaneous

Component board, 8 ohm loudspeaker, wire, solder, etc.

10 Watt VMOS Amplifier

Figure 8 shows the circuit diagram of a good quality 10 watt RMS amplifier using a quasi complementary, Class B VMOS output stage.

As in the previous designs, a source follower transistor (Tr3 in this case) is used to drive the load when the output is positive going, and bootstrapping (provided by D1 and C6) is used to give good efficiency on positive going outputs. When the output is negative going the load is driven by Tr3 which is used in the common source mode, and as in the previous circuit, this gives good efficiency on negative going outputs as well. Tr3 is preceded by an operational amplifier device which has its non-inverting input connected to the output of the amplifier (Tr3 and Tr4 sources). Allowing for the fact that the signal is inverted through Tr4, this gives 100% negative feedback over Tr4 and IC1. This gives unity voltage gain from the inverting input of IC1 to the output of the amplifier, so that a reasonably good complement to Tr3 is obtained. There is a slight mismatch due to the fact that Tr4 and IC1 give almost exactly unity, whereas the voltage gain of Tr3 (as explained earlier) will actually be significantly less than unity. This is of no real consequence though, since the circuit uses a considerable amount of overall negative feedback, and this smoothes out the mismatch and ensures good distortion performance. IC1 is a type which has a P MOS input stage, and it therefore provides an extremely high input impedance which is comparable to that of a VMOS device.

The driver stage uses a VMOS device in the common source mode (Tr2), and this has R9 as its main collector load and R10 to give the appropriate quiescent bias current through the

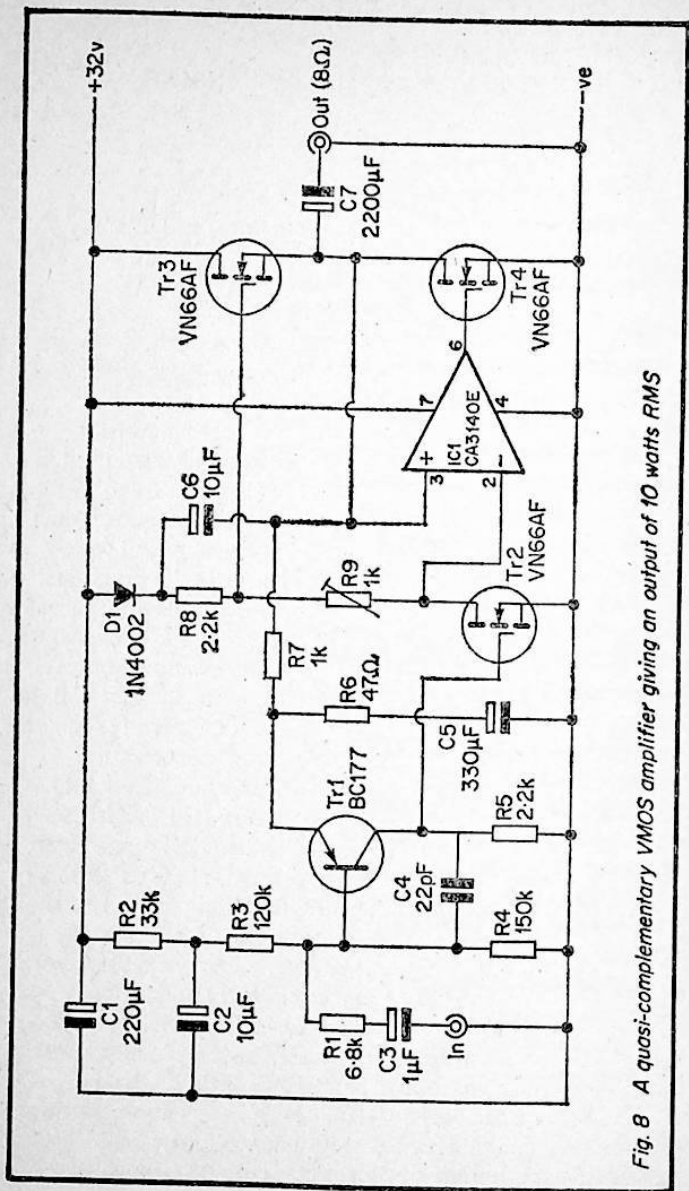


Fig. 8 A quasi-complementary VMOS amplifier giving an output of 10 watts RMS

output transistors. R10 should be set at almost minimum resistance initially, so that a high bias current through the output transistors (and the possibility of damaging these devices in consequence) is avoided. R10 is then adjusted for increased resistance until a current of about 30mA is drawn by the amplifier.

Tr1 is used as the common emitter input stage, and this is directly coupled to Tr2 with R6 forming the collector load for Tr1. Although in circuits where a bipolar driver stage is used in this basic configuration the circuit will operate without a load resistor for the input stage (the base-emitter junction of the driver transistor providing a high but useable resistance), in this circuit the resistor is essential. The input impedance of a VMOS device is far too high to provide a suitable collector load resistance for a bipolar transistor. In fact the input impedance of a VMOS device is so high that without R6, Tr1 would tend to charge up the input capacitance of Tr2, and it would then stay charged as there is no external discharge path, and the gate to source resistance of Tr2 would be too high to provide a suitable path. Thus Tr2 would switch on, the output voltage of the amplifier would drop to virtually zero, and the circuit would latch in this state. This is something that should be borne in mind when designing the drive circuit for a VMOS device.

R8 couples the emitter to the output of the amplifier so that 100% negative feedback and unity voltage gain are obtained at DC. R2 to R4 are used to bias the amplifier, and R2 plus C2 prevents hum or other noise on the supply lines from being coupled into the input of the amplifier by way of the bias circuit. R7 and C5 remove some of the feedback applied to the amplifier at audio frequencies, giving a voltage gain of roughly twenty times (26dB). The gain of the amplifier can be varied somewhat by changing the value of R7, the voltage gain being roughly equal to R8 divided by R7. C4 rolls off the high frequency response of the circuit to a limited extent, and this aids stability. With a well designed circuit layout it may well be found that C4 is unnecessary, although it might still be advisable to include it in order to help prevent problems with radio frequency interference breaking through to the output.

With a 32 volt supply the circuit will give an output of typically 10 watts RMS into an 8 ohm load, and the total harmonic distortion is only about 0.1% at most output powers, but does rise significantly just before the onset of clipping. About 450mV RMS is needed at the input to produce maximum output power, and the input impedance of the circuit is approximately 60k.

The supply voltage should not be allowed to exceed 36 volts, which is the absolute maximum supply rating of the CA3140 device used in the IC1 position. The supply current under maximum output conditions is about 600mA, and the supply should be able to supply this current while giving an output potential of 32 volts if the maximum output power of the circuit is to be realised. In practice this means that either a stabilised supply must be used, or a mains transformer having a very generous secondary current rating should be used in a non-stabilised circuit (a 24 volt 1.5 ampere type followed by a bridge rectifier and 3,300 μF of smoothing is suitable).

Tr3 and Tr4 both have to dissipate a substantial amount of power under high output conditions, and they should be mounted on large heatsinks.

Components: Quasi-Complementary 10 Watt Amplifier (Figure 8)

Resistors (all 1/3 watt 5% except R10)

R1	6.8k	R6	47 ohms
R2	33k	R7	1k
R3	120k	R8	2.2k
R4	150k	R9	1k 0.25W preset
R5	2.2k		

Capacitors

C1	220 μF 40V	C5	330 μF 25V
C2	10 μF 40V	C6	10 μF 25V
C3	1 μF 40V	C7	2200 μF 40V
C4	22pF ceramic plate		

Semiconductors

IC1	CA3140E (or CA3140T)		
Tr1	BC177	Tr3	VN66AF
Tr2	VN66AF	Tr4	VN66AF

D1 1N4002

Miscellaneous

Component panel, 8 ohm speaker capable of handling at least 10 watts RMS, wire, solder, etc.

Alternative 10W design

There are, of course, other methods of producing a quasi-complementary VMOS output stage, and the circuit of Figure 9 demonstrates one such method. This uses a system which is very similar to that employed in early bipolar quasi-complementary designs. These often used a Darlington pair emitter follower stage in the upper section of the output stage, with two common emitter amplifiers having 100% negative feedback being used in the lower section. The Darlington pair emitter follower give no phase inversion and approximately unity voltage gain. The two emitter follower amplifiers each generate a phase inversion, giving no overall phase change. Due to the high level of feedback they produce only about unity voltage gain. This method therefore gave quite good and symmetrical results, although these days true complementary arrangements are possible and normally used.

In this circuit Tr4 is used to effectively replace the Darlington pair emitter follower stage, Tr5-6 is the common emitter (or common source to be more accurate) output transistor, and Tr3 is the common emitter complementary device for Tr5-6. R6 acts as the collector load for Tr3, and the 100% negative feedback is obtained merely by taking the emitter of Tr3 to the output of the amplifier.

One drawback of this arrangement in a VMOS design is that the threshold voltage of a VMOS device is higher than that of a bipolar transistor, and a fairly substantial gate to source voltage is needed in order to drive a VMOS device hard into the on state. The gate voltage for Tr5-6 is obtained from the output of the amplifier by way of Tr3. In order to take the output of the amplifier fully negative it is necessary for Tr5-6 to have a fairly high gate to source voltage so that it is biased hard on, but as the output of the amplifier swings more negative, so the maximum gate to source voltage that Tr5-6 can achieve reduces. This tends to give only a fairly

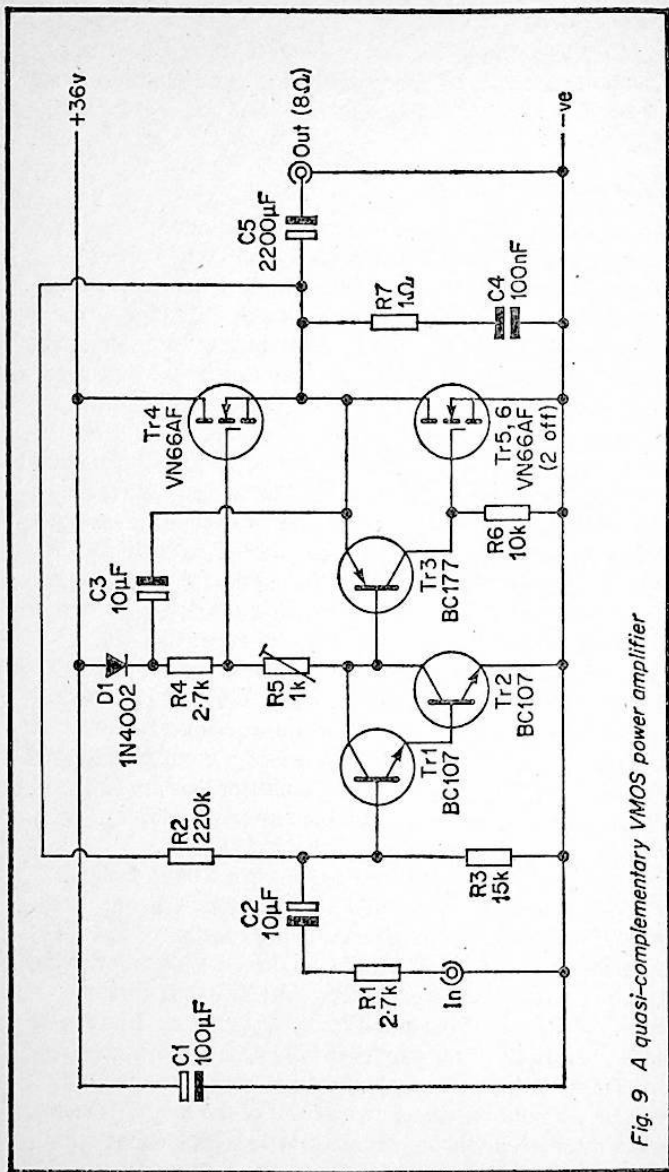


Fig. 9 A quasi-complementary VMOS power amplifier

low level of efficiency.

One way of alleviating this problem to some extent, and the method used in this design, is to use two VMOS devices connected in parallel in the common source section of the output stage. This reduces the gate bias potential needed for a given output current, and thus gives improved efficiency by enabling the output to swing closer to the negative supply rail voltage.

The rest of the circuit is quite straight forward with Tr1 and Tr2 being used as the basis of a Darlington pair common emitter driver stage. R5 is used to set the appropriate current through the output transistors under quiescent conditions (as described in the previous section of this book for the circuit of Figure 8), and D1 plus C3 provide bootstrapping so that good efficiency is obtained from Tr4. R2 and R3 bias the output of the amplifier to slightly more than half the supply potential, the output being offset towards the positive supply to allow for the fact that the output can swing closer to the positive rail than it can to the negative one. R7 and C4 form a Zobel network, and these aid the stability of the amplifier. No other stabilisation components are required, and it may well be found that R7 and C4 are unnecessary.

Using a supply voltage of about 36 volts or so the circuit achieves a similar performance to that of Figure 8, but has a slightly lower gain and input impedance. A little over 500mV RMS is needed at the input in order to give maximum output, and the input impedance is approximately 15k.

*Components: Quasi-Complementary VMOS Amplifier
(Figure 9)*

Resistors (all 1/3 watt 5% except R5)

R1	2.7k	R5	1k 0.25 watt preset
R2	220k	R6	10k
R3	15k	R7	1 ohm
R4	2.7k		

Capacitors

C1	100 μ F 40V	C4	100nF plastic foil
C2	10 μ F 25V	C5	2200 μ F 40V
C3	10 μ F 25V		

Semiconductors

Tr1 BC107

Tr2 BC107

Tr3 BC177

D1 1N4002

Tr4 VN66AF

Tr5 VN66AF

Tr6 VN66AF

Miscellaneous

Component panel, 8 ohm speaker capable of handling at least 10 watts RMS, wire, solder, etc.

20W MOS Amplifier

The Hitachi power MOSFETs were primarily developed for use in high quality audio power amplifiers, and have the same advantages over bipolar devices as VMOS transistors. Although VMOS transistors and power MOSFETs have similar characteristics, their structures are very different. Power MOSFETs are horizontal current flow devices (as are ordinary low power field effect devices) and have the cross section shown in Figure 10.

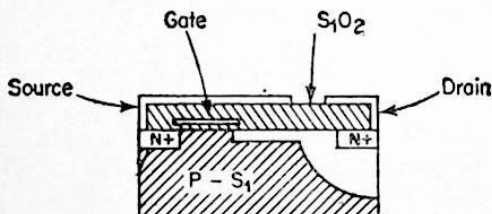


Fig. 10 Cross section of an n-channel power MOSFET

Six types of power MOSFET are available to the amateur user at the time of writing; three P channel devices and matching complementary N channel devices. The principle difference between these three sets of transistors is their maximum drain to source voltage, as can be seen from the table provided below.

<i>Device</i>	<i>Channel</i>	<i>Maximum Drain-Source Voltage</i>
2SK133	N	120V
2SJ48	P	120V
2SK134	N	140V
2SJ49	P	140V
2SK135	N	160V
2SJ50	P	160V

All these devices are contained in a standard metal TO-3 encapsulation. However, this does not use the same pinout configuration as bipolar devices. Whereas the case of a TO-3 bipolar transistor connects to the collector terminal of the device (which is equivalent to the drain of a FET) the case of a TO-3 power FET connects to the source terminal of the device (the equivalent of a bipolar transistor's emitter). This is a useful feature when these devices are used in Class B audio output stages as source followers, as the two sources then connect together. The transistors can therefore be mounted on a single heatsink without the need to insulate them from the heatsinks, although the heatsink must, of course, be electrically isolated from the rest of the circuit.

All six devices have a maximum power rating of 100 watts and a maximum drain current rating of 7 amperes! The "on" resistance is 1.71 ohms (maximum), or about 1 ohm (typical).

One advantage of power MOSFETs over VMOS devices is the higher gain of the former. Most VMOS devices have a forward transconductance of about 250m Ω as opposed to a typical figure of 1,000m Ω for a power MOSFET (although a few N channel, high power VMOS devices do have typical forward transconductance figures in excess of 1,000m Ω). This helps to give reasonably good efficiency when the device is used in the source follower mode. This is also aided by the fact that the gate threshold voltage of a power MOSFET is lower than that of a VMOS device, or a silicon bipolar transistor come to that. This can be seen by referring to the graph of Figure 11 which shows the typical transfer characteristic of a power MOSFET.

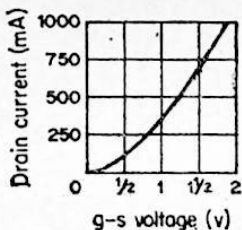


Fig. 11 Typical transfer characteristics of a power MOSFET

20W MOSFET Amplifier

Power MOSFETs are suitable for use in the highest quality audio power amplifiers. The Hitachi Power MOSFET application note gives details of 50W and 100W designs offering typical distortion levels of around 0.005%, and only about 0.01% total harmonic distortion at full output and a frequency of 100kHz! Performance of this order would be difficult (probably impossible) to obtain using bipolar devices. At the time of writing there are a few power MOSFET amplifier kits available offering similar levels of performance.

Like VMOS devices, power MOSFETs have the advantage of very fast operating speeds and consequent consistency of performance to beyond the limit of the audio frequency spectrum. They are also free of the phase lags that can cause problems if high levels of negative feedback are used at high frequencies. They also have an extremely high input resistance, of course, but the input impedance does drop significantly at higher audio frequencies due to the relatively high input capacitance of about 500 to 600pF. This is about ten times or so higher than the input capacitance of a typical VMOS transistor, although high power VMOS devices do have a comparable input capacitance. Anyway, a current of only about 10mA in the driver stage is sufficient to give consistent results to beyond the upper limit of the audio range of frequencies, and the higher input capacitance is of no real

consequence in audio frequency applications.

The circuit of Figure 12 is provided for those who would like to try out a simple amplifier utilizing power MOSFETs in the output stage. This uses a basic configuration which is similar to that employed in some of the designs covered earlier in this book, with a common emitter input transistor (Tr1) directly driving a common source VMOS driver device (Tr2) which in turn directly drives the complementary common source output transistors (Tr3 and Tr4). R5 provides 100% negative feedback over the amplifier at DC so that R1 to R3 can be used to bias the output to the appropriate potential. C6 and R4 decouple some of the feedback at audio frequencies and give a voltage gain of about 20 times (26dB). This gives the circuit an input sensitivity of approximately 625mV RMS into 70k for an output power of 20 watts RMS.

R8 is used to set the appropriate quiescent current through the output transistors, and about 80 to 100mA is a suitable level. Power MOSFETs are negative coefficient devices, and so no temperature compensation circuitry is required. R9 and C4 form a low pass filter at the input of the circuit, and this helps to combat problems of radio frequency interference. C5 rolls off the response of the circuit slightly at high frequencies, aiding stability and also helping to avoid radio frequency breakthrough. C3 and C8 are the input and output DC blocking capacitors respectively.

Using a 50 volt (loaded) supply and an 8 ohm speaker this circuit provides an output power of 20 watts RMS with ease. An output power of about 15 watts RMS can be achieved using a loaded supply voltage of about 40 volts or so, and some 30 watts RMS can be achieved using a 60 volt loaded supply. Although the circuit would not qualify for the super-fi category by current standards, it does achieve a remarkable level of performance for a design of such simplicity (it does, after all, only use four transistors)! The total harmonic distortion is typically well below 0.1% at most output powers and frequencies, although it does rise slightly at high and low output powers, and at high frequencies (as one would expect).

Although the BC177 device used in the Tr1 position does have a maximum emitter to collector voltage rating of just 45 volts, it is safe to use this device in this circuit with a supply

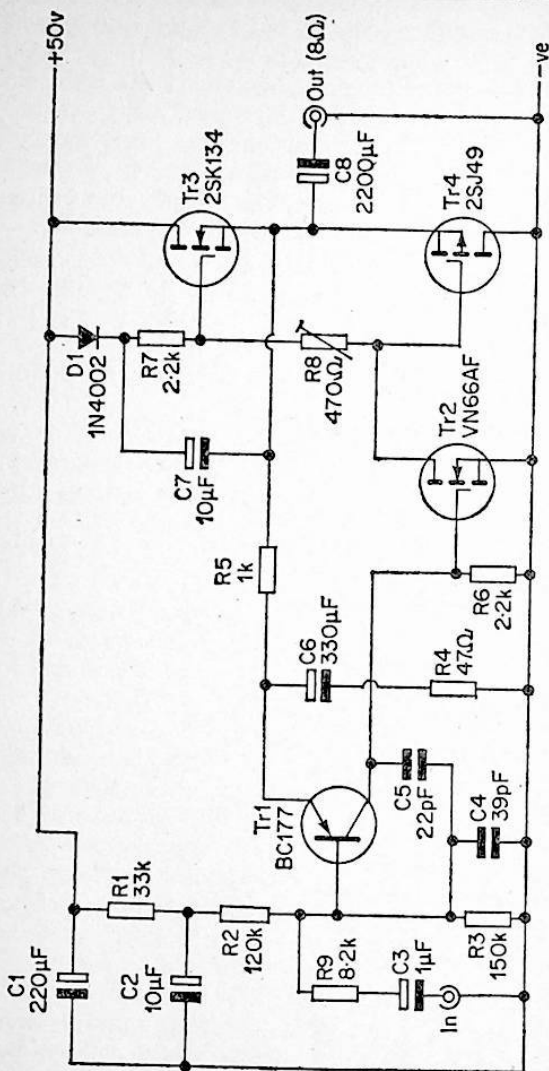


Fig. 12 The circuit diagram of a 20 watt RMS MOS amplifier

voltage of well in excess of this figure. This is simply because the peak voltage across the emitter and collector terminals of Tr1 is little more than half the supply potential. However, if preferred, a high voltage, high gain, low noise transistor such as the 2SA872A can be used in the Tr1 position. A 2SK134 and a 2SJ49 were used as the output transistors in the prototype amplifier, but the 2SK133/2SJ48 and 2SK135/2SJ50 devices are also suitable for use in the circuit.

The unit requires a substantial supply current when operating at high output powers, the approximate figures being 690mA at 15 watts RMS, 800mA at 20 watts RMS, and 980mA at an output power of 30 watts RMS.

Components: 20 Watt MOS Amplifier (Figure 12)

Resistors (all 1/3 watt 5% except R7 and R8)

R1	33k	R6	2.2k
R2	120k	R7	2.2k 1 watt
R3	150k	R8	470 ohms 0.25 watt preset
R4	47 ohms	R9	8.2k
R5	1k		

Capacitors

C1	220 μ F 63V	C5	22pf ceramic plate
C2	10 μ F 63V	C6	330 μ F 40V
C3	1 μ F 40V	C7	10 μ F 40V
C4	39pf ceramic plate	C8	2200 μ F 63V

Semiconductors

Tr1	BC177	Tr3	2SK134 (see text)
Tr2	VN66AF	Tr4	2SJ49 (see text)
D1	1N4002		

Miscellaneous

Component panel, loudspeaker capable of handling at least 20 watts RMS, heatsinks, wire, solder, etc.

Simple Class A Amplifier

With the decline of the thermionic valve, class A power amplifiers have dropped considerably in popularity and have become something of a rarity. Class A designs are generally more simple than Class B types, but have the disadvantage of being very inefficient. For example, a power supply and set of

output transistors that would comfortably give an output power of (say) 25 watts RMS in a Class B design would probably only give about 8 watts in a Class A circuit. In lower power designs, which are often battery powered, the low efficiency of Class A designs has the disadvantage of reduced battery life and hence increased running costs.

Class A amplifiers are still used to some extent, and they are still to be found in very low power applications where the low efficiency is not of great importance due to the fact that little heat will be generated in the output stage, and only a modest supply current is needed despite the poor efficiency.

VMOS transistors are well suited to such designs, and Figure 13 shows the circuit diagram of a simple class amplifier using a VN10KM VMOS device as the output transistor. With the relatively high load impedance and low powers involved here the VN10KM is more than adequate, but the circuit will, of course, also operate properly using medium power devices such as the VN66AF or VN67AF if one of these should be to hand or more readily available.

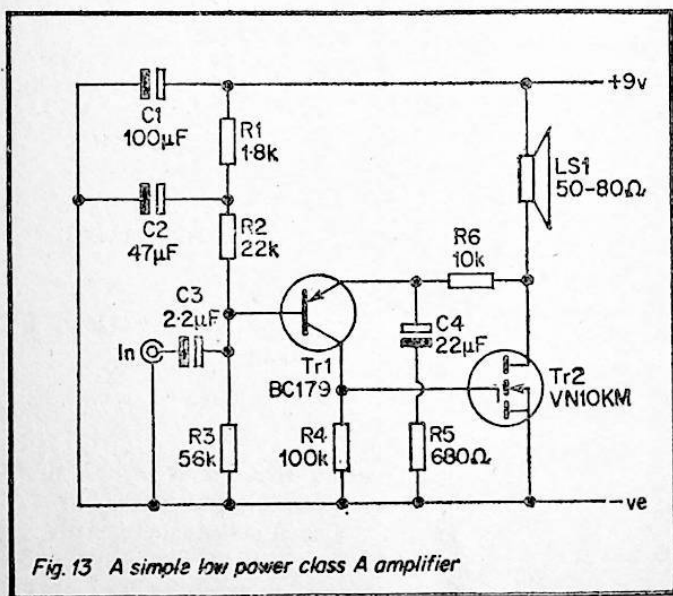


Fig. 13 A simple low power class A amplifier

The circuit uses Tr1 as a common emitter driver stage which directly drives Tr2; the common source output stage. The circuit is, in fact, very much like the input and driver stages of some of the designs covered earlier, but there is no push pull common source (or emitter) output stage with the loudspeaker instead being used as the drain load for Tr2. R1 to R3 bias the circuit so that there is a quiescent current of about 25mA through LS1 and Tr2. The input signal varies this current either side of the quiescent figure, thus driving LS1. The output current varies from 0 to about 50mA at maximum output, and the average current consumption therefore remains constant at 25mA, and does not rise at high output levels as in the case of a class B amplifier. The output power is only about 23mW RMS, but this is adequate for a number of applications (small radios, intercoms, etc.). The circuit gives reasonably good reproduction quality; the main limitation on the output quality being the distortion and limited frequency response of the miniature high impedance loudspeaker used.

An input of only about 80mV RMS is needed in order to produce maximum output, but this figure can be changed, within reason, by altering the value of R5. Changes in the value of this component cause an inversely proportional change in the input sensitivity of the circuit.

Do not use a low impedance loudspeaker in this circuit as this would result in a high current flow through Tr2 and the loudspeaker, probably resulting in the destruction of one or both of these.

Components: Low Power Class A Amplifier (Figure 13)

Resistors (all 1/3 watt 5%)

R1	1.8k	R4	100k
R2	22k	R5	680 ohms
R3	56k	R6	10k

Capacitors

C1	100 μ F 10V	C3	2.2 μ F 10V
C2	47 μ F 10V	C4	22 μ F 10V

Semiconductors

Tr1	BC179	Tr2	VN10KM (or VN66AF VN67AF, etc.)
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Miscellaneous

Component board, high impedance (50 to 80 ohm) loudspeaker, wire, solder, etc.

Alternative Version

One slight problem with the circuit of Figure 13 is that it results in a standing current through the loudspeaker so that the cone of the speaker does not move backwards and forwards about its normal resting position, but is always offset one way or the other. This can result in a reduction in performance from the loudspeaker.

This can be avoided using the modified circuit shown in Figure 14. Here Tr2 has a load resistor (R7) and the voltage swing at Tr2 drain is coupled to the loudspeaker by C5. The bias circuit (R1 to R3) is modified slightly in order to give a quiescent voltage at Tr2's drain terminal of about half the supply potential, and this gives the same nominal quiescent bias current through Tr2 as in the original circuit.

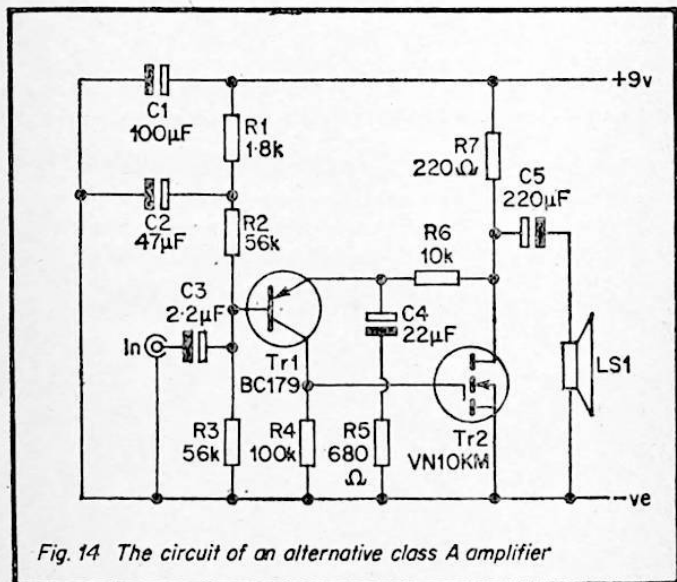


Fig. 14 The circuit of an alternative class A amplifier

The circuit of Figure 14 gives a similar level of performance to that of Figure 13, but the output power is slightly less since, in effect, Tr2 is driving R7 as well as LS1, and a certain amount of power is wasted in R7.

Components: Alternative Class A Amplifier (Figure 14)

All components required for the original circuit, plus 220 ohm resistor (R7) and 220 μF 10V capacitor C5. Also, R2 becomes 56k.

7W Class A Design

Class A designs are not often used when a higher power amplifier is required, but they are preferred by some people because no crossover distortion is produced by Class A designs. We have already seen how crossover distortion is produced by a Class B amplifier, and how it can be reduced by applying a quiescent bias current through the output transistors. However, this bias only reduces crossover distortion, and does not completely eliminate it.

This is simply because semiconductor devices tend to have a relatively low level of gain at and just above the threshold voltage. This is something which is common to bipolar, VMOS, and power MOSFET devices. It is an effect which can be clearly seen in both the VMOS transfer characteristic of Figure 4 and the power MOSFET transfer characteristic of Figure 11. In both cases the gain gradually increases until a substantial drain current is achieved, after which a virtually linear transfer is obtained.

Unfortunately, crossover distortion tends to increase as output power decreases (since at low powers the output devices are operating largely or totally over a very non-linear part of the transfer characteristic), and distortion tends to be most noticeable at low volume levels.

Although modern, good quality Class B designs produce very low levels of crossover distortion, some people nevertheless prefer the "sound" of Class A circuits. These do not produce crossover distortion because there is a heavy quiescent current through the output transistor, and under low signal conditions it is used over a reasonably linear part of the output transistor's transfer characteristic. It is only when almost fully driven that drain current drops down to a fairly low level during signal peaks on one set of half cycles, causing a rise in distortion.

This rise in distortion is of relatively little importance since distortion tends to be less noticeable at high volume levels. Also, if used with reasonably efficient speakers, it is likely that the amplifier would never be driven hard enough to bring the output

transistor into the non-linear operating region, and the rise in distortion would not actually occur in practice.

The circuit diagram of Figure 15 is for a good quality Class A amplifier which uses a VMOS driver stage, power MOSFETs in the output stage, and provides an output power of about 7 watts RMS. The basic configuration of the amplifier is the same as that used in several of the circuits described earlier with Tr1 acting as a common emitter input stage, Tr2 as a common source driver stage, and Tr4 as a source follower output stage. Although at a first glance it might appear as though a complementary Class B output stage is used, closer examination of the circuit will reveal that only Tr4 is driven by Tr2, and Tr5 is actually used in a constant current generator which forms the source load for Tr4.

A load resistor or output transformer might seem to be better choices as the load for Tr4, but a constant current source is probably the best choice in practice for the following reasons. Although an output transformer may seem to offer good results, it would inevitably have a significant effect on the frequency response and distortion of the circuit (it is possible to counteract this to some extent by applying negative feedback over the entire amplifier including the output transformer). Another problem with an output transformer is that it would be designed to give a minimum loss of quality rather than optimum signal transfer, and could easily introduce losses in the region of 30 to 40%. Thus, although in theory a transformer offers good efficiency, in practice things would probably be quite different. However, the main problem with using a transformer is simply that it would almost certainly be impossible to obtain a suitable component.

A low value resistor could be used as the source load for Tr4, but as the output of the amplifier swung positive, Tr4 would be driving both the speaker and the load resistor, causing a large current to flow through the load resistor. The current through the resistor would serve no useful purpose, and would merely give the circuit low efficiency.

The same problem does not exist with the constant current load, since this will not draw an increased level of current as the output swings positive. The current drawn by the amplifier does increase though, as a current is supplied to the speaker as well as to the constant current load. As the output swings

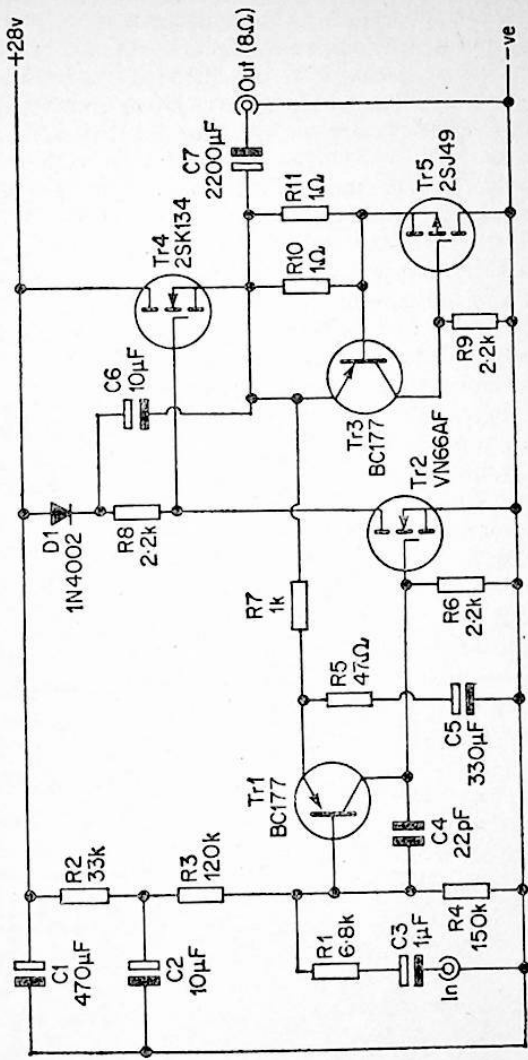


Fig. 15 The circuit diagram of a 7 watt RMS class A VMOS amplifier

negative the current flow in the speaker circuit is reversed, and the output current then flows through the constant current load and not through Tr4. The output current is then provided by the output coupling capacitor (C7) as it discharges, as is the case with an ordinary Class B output stage. When the amplifier is fully driven C7 provides all the current for the constant current generator, and Tr4 switches off, and draws no current from the supply. Thus the supply current varies from zero to double the quiescent current under full load conditions, giving an average current consumption equal to the quiescent current consumption. The average current consumption of the circuit therefore remains constant, like more conventional Class A amplifier configurations.

Of course, even under full output conditions there is a certain amount of power wasted in the constant current load, and the circuit is not very efficient, but this is something one has to accept with any true Class A circuit. Optimum efficiency would be obtained by using the speaker itself as the source load for Tr4, but placing a high standing current through loudspeakers is not to be recommended, and could result in eventual displacement of the speech coil. In any event, it would be unlikely to produce optimum results with regard to quality. Most good quality speakers incorporate two or more drivers and a crossover network, and it is possible that the latter could give rise to problems if the direct drive method were to be used.

The peak output current that the circuit can provide is obviously equal to the quiescent current set by the constant current load. This current is set by the parallel resistance of R10 and R11 at approximately 1.3 amperes, and with a 8 ohm load this gives a peak output power of about 14 watts, and an RMS output power of about 7 watts. The distortion is typically below 0.1% at all power levels up to the point just below the clipping level.

Components: 7 Watt Class A Amplifier (Figure 15)

Resistors (all 1/3 watt 5% except R10 and R11)

R1	6.8k	R5	47 ohms
R2	33k	R6	2.2k
R3	120k	R7	1k
R4	150k	R8	2.2k

R9 2.2k
R10 1 ohm 1 watt

R11 1 ohm 1 watt

Capacitors

C1 470 μ F 40V
C2 10 μ F 40V
C3 1 μ F 40V
C4 22pf ceramic plate

C5 330 μ F 25V
C6 10 μ F 25V
C7 2200 μ F 25V

Semiconductors

Tr1 BC177
Tr2 VN66AF
Tr3 BC177
D1 1N4002

Tr4 2SK134
Tr5 2SJ49

Miscellaneous

Component board, 8 ohm loudspeaker capable of handling at least 7 watts RMS, heatsinks, wire, solder, etc.

Constant Current Generator

The constant current generator used in this circuit uses what is basically a well known configuration, but a power MOSFET is used where a bipolar device would normally be employed. A power FET enables a high current constant current generator to be produced using very few components due to its minimal drive current requirement. One slight drawback of using a power FET in this application is that the minimum voltage which the circuit will provide at full output may be slightly higher than in a bipolar design, although this depends to some extent on the level of output current involved and the particular designs used, and in most cases a power FET will be only marginally less efficient.

Operation of this constant current generator is quite straight forward. Tr5 is used in the source follower mode, and is biased hard into conduction by R9. This causes a heavy current to flow through the resistance in the source circuit provided by R10 and R11 in parallel (these provide a resistance of 0.5 ohms). Two 1 ohm resistances were used here because they were to hand, and are readily available, but a 0.47 ohm component having a power rating of 2 watts or more is also suitable.

The current flowing through R10 and R11 causes a voltage

to be developed across these components, and it will be of sufficient magnitude to switch on Tr3. Tr3 then tends to reduce the gate to source voltage of Tr5, preventing the voltage across R10 and R11 from rising far above the threshold voltage of Tr3. As Tr3 is a silicon device, the circuit stabilises with a potential of roughly 0.65 volts across R10 and R11, and from Ohm's Law it can be seen that this gives an output current of about 1.3 amperes ($0.65V \div 0.5 \text{ ohms} = 1.3 \text{ amperes}$), provided there is a sufficiently low load impedance for the circuit, such as the source follower circuit in this circuit.

Tr4 and Tr5 dissipate about 18 watts each, and must therefore be mounted on substantial heatsinks. Note that in this circuit the source terminals of the power MOSFETs do not connect together, and they must either be fitted on separate heatsinks, or at least one of them must be properly insulated from the heatsink if a common one is used. A very high output current will flow if the two source terminals should become short circuited, and this could easily destroy Tr4, Tr5 and some of the power supply components!

CHAPTER 3

SOUND GENERATOR CIRCUITS

The low drive current requirement of VMOS devices makes them ideal for use in sound generator circuits, particularly where a modulated, high power signal is required. The high gain of a VMOS device enables it to directly interface a low power signal source to a low impedance speaker driven at several watts.

In this chapter a few sound generator circuits will be described, ranging from a low power monotone type to high power modulated tone circuits. These are suitable for use in burglar alarm systems and in other alarm applications.

Low Power Alarm

The circuit diagram of Figure 16 is for a simple single tone alarm generator which has an output power of about 300mW. Tr2 is biased by R4 and R5 to operate as a Class A common source stage having high impedance loudspeaker LS1 as its collector load. Tr1 is used as a common emitter stage having

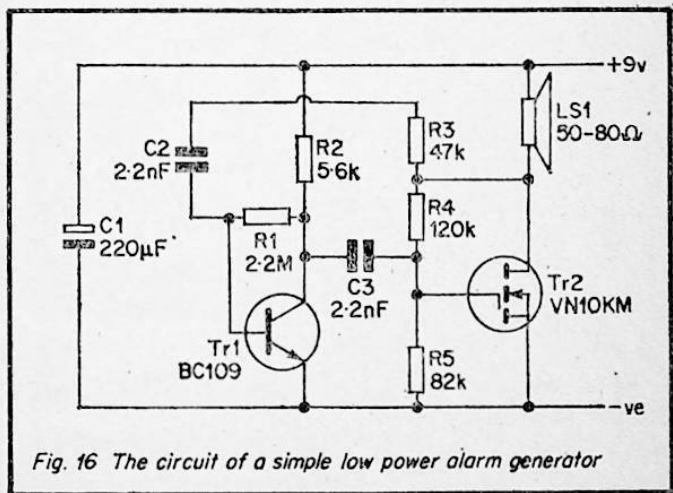


Fig. 16 The circuit of a simple low power alarm generator

R2 as its collector load, and base biasing provided by R1. C3 couples the output from the collector of Tr1 to the input (gate terminal) of Tr2.

Both Tr1 and Tr2 provide an inversion of the signal, so that Tr1's base is in phase with Tr2's collector. Positive feedback is therefore provided through C2 and R3, and the gain of the circuit is large enough to produce strong oscillation. This gives a roughly squarewave signal at the drain terminal of Tr2, with this device being switched between the hard on and hard off states. This results in quite substantial pulses of current (around 140mA peak) being fed through the loudspeaker, and a quite loud audio tone is produced. The operating frequency is approximately 1kHz, but can be changed by altering the values of C2 and C3. These should have the same value, and changes in value produce an inversely proportional change in operating frequency

Dissipation in Tr2 is quite low and so a VN10KM device is adequate in the Tr2 position (although larger devices such as the VN66AF and VN67AF are also suitable). The circuit should not be used to drive a low impedance speaker as there would then be quite a high dissipation in Tr2, probably leading to it overheating and being destroyed.

The current consumption of the circuit is approximately 70mA.

Components: Simple Low Power Alarm Generator (Figure 16)

Resistors (all 1/3 watt 5% (10% over 1M))

R1	2.2M	R4	120k
R2	5.6k	R5	82k
R3	47k		

Capacitors

C1	220 μ F 10V	C3	2.2nF plastic foil
C2	2.2nF plastic foil		

Semiconductors

Tr1	BC109	Tr2	VN10KM (or VN66AF, VN67AF, etc)
-----	-------	-----	---------------------------------

Miscellaneous

Component board, high impedance (50 to 80 ohm) loudspeaker, wire, solder, etc.

High Power Version

Where a higher output power is required the circuit of Figure 17 can be used. This is basically similar to the design shown in Figure 16, but uses a higher supply voltage and lower impedance speaker in order to provide increased output power. The actual output power is about 3.5 watts using an 8 ohm loudspeaker, or approximately 2 watts using a 15 ohm speaker. In either case the volume of the sound generated by the unit is decidedly loud provided a reasonably efficient loudspeaker is used, and the unit is suitable for use in applications such as burglar alarms, where high volume is obviously essential.

As was the case with the previous circuit, the operating frequency is about 1kHz, but can be altered by changing the values of C2 and C3. Also as before, the operating frequency is inversely proportional to the value of these components.

It is not advisable to use the circuit with a speaker having an impedance of less than 8 ohms as the efficiency would not be very good, resulting in high dissipation in Tr2. Also, the maximum current rating of Tr2 could be exceeded. There is a significant amount of power dissipated in Tr2, and it should be fitted with at least a small bolt-on finned heatsink. The average

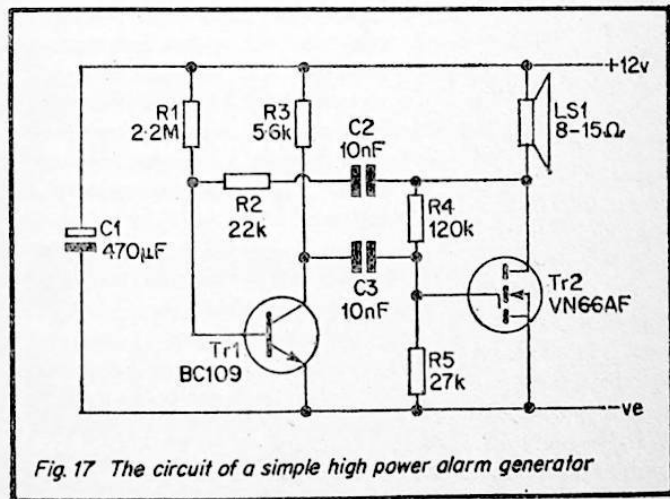


Fig. 17 The circuit of a simple high power alarm generator

supply current using an 8 ohm loudspeaker is approximately 600mA, or about 350mA or so if a 15 ohm loudspeaker is used.

Components: Simple High Power Alarm Generator (Figure 17)

Resistors (all 1/3 watt 5% (10% over 1M))

R1	2.2M	R4	120k
R2	22k	R5	27k
R3	5.6k		

Semiconductors

Tr1	BC109	Tr2	VN66AF
-----	-------	-----	--------

Capacitors

C1	470 μ F 16V	C3	10nF plastic foil
C2	10nF plastic foil		

Miscellaneous

Component board, 8 to 15 ohm impedance speaker (see text), wire, solder, etc.

Pulsed Tone Generator

Simple alarm circuits of the type described so far are quite effective, but not as effective as more complicated circuits that give a modulated output of some kind. This is merely because a single output tone is rather monotonous and more easily masked by other sounds than a signal which is constantly changing. The more complex sound of a modulated alarm also tends to be more attention catching than a straight forward and relatively uninteresting single tone.

One method of modulating a tone to produce a more effective signal, and probably the most simple method, is to amplitude modulate the tone using a squarewave modulating signal. In effect, this is just switching the tone on and off so as to make it a little less monotonous, and in practice this method is quite effective.

The circuit diagram of a simple low power pulsed tone generator is shown in Figure 18. This is based on a CMOS 4001 quad 2 input NAND gate, but three of the gates (1, 2 and 4) have their inputs connected in parallel so that they actually operate as simple inverters.

Gates 1 and 2 are used in a straight forward CMOS astable multivibrator circuit, and the values of R1 and C2 have been

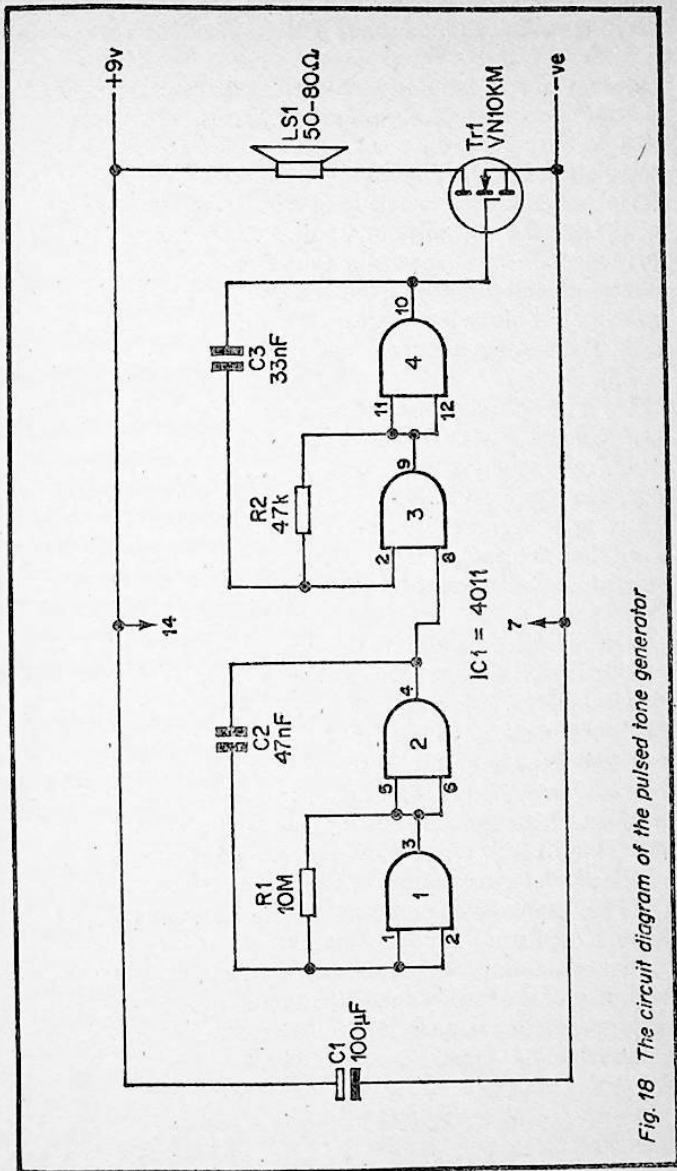


Fig. 18 The circuit diagram of the pulsed tone generator

selected to give an operating frequency of about 1 Hertz. Gates 3 and 4 are also used in a CMOS astable circuit, but the values of R2 and C3 give this astable a much higher operating frequency of approximately 800 Hertz. However, one input of gate 3 is driven from the output of Gate 2. When gate 2 takes this input low, the output of gate 3 will go high and the logic state at the other input of gate 3 becomes irrelevant. The second astable is then unable to operate. When the output of gate 2 takes the controlled input of gate 3 high, a normal inverter action is obtained from the other input of gate 3 to the output, and this astable then functions normally. Thus, as the output of the low frequency astable alternates between the high and low states, the higher frequency oscillator is pulsed on and off.

Tr1 is used as a common source amplifier which drives loudspeaker LS1 with the audio tone when the controlled astable is functioning. The output of gate 4 goes low when the tone generator is muted, cutting off Tr1. This is essential, as otherwise a high current would probably flow through Tr1 and LS1 during the periods between bursts of output tone, giving greatly increased current consumption and dissipation in LS1 and Tr1.

VMOS devices are ideal for use with CMOS circuits which must drive high current loads, incidentally. CMOS devices produce quite a large output voltage swing; virtually equal to the supply potential in fact. Such a voltage swing is only produced if the output is lightly loaded though, since CMOS logic ICs have a fairly high output impedance. VMOS devices require a fairly large drive voltage, but very little current, and CMOS logic ICs therefore provide an ideal drive signal for them.

The current consumption of this circuit averages about 70mA while the tone is being produced, but is practically zero when the tone oscillator is muted. This gives an overall average current consumption of only about 35mA. If desired, the frequency of the tone and modulation oscillators can be altered, and the operating frequencies of these are inversely proportional to the values of C3 and C2 respectively.

Components: Pulsed Tone Generator (Figure 18)

Resistors: (all 1/3 watt 5% (10% over 1M))

R1 10M R2 47k

Capacitors

C1 100 μ F 16V C3 33nF plastic foil

C2 47nF plastic foil

Semiconductors

IC1 4011

Tr1 VN10KM (or VN66AF, VN67AF, etc.)

Miscellaneous

Component board, high impedance (50 to 80 ohm) loudspeaker,

Wire, solder, etc.

High Output Version

If a pulsed tone generator having a higher output power is required, this can be achieved by increasing the supply potential to 12 volts, using an 8 or 15 ohm loudspeaker, and changing Tr1 to a VN66AF device fitted with a small finned heatsink. This gives an increased overall average current consumption of about 300mA using an 8 ohm loudspeaker, or 175mA using a 15 ohm impedance component.

FM Tone Generator

Another very effective form of alarm generator is the frequency modulated type. Here the frequency of the output tone is varied in some way. Using a squarewave modulating signal for example, the tone is switched between two pitches at a rate equal to the frequency of the modulation oscillator. Using a triangular modulation signal causes the pitch of the output signal to be smoothly varied up and down in pitch at the modulation oscillator's operating frequency.

The circuit of Figure 19 shows how two CMOS astable circuits can be connected to produce a frequency modulated tone generator. Gates 1 and 2 are used in the modulation astable, and timing components R1 and C2 give an operating frequency of roughly 5 Hertz. Gates 3 and 4 are used in the tone generator astable, and R3 plus C3 give a nominal operating frequency of about 800 Hertz.

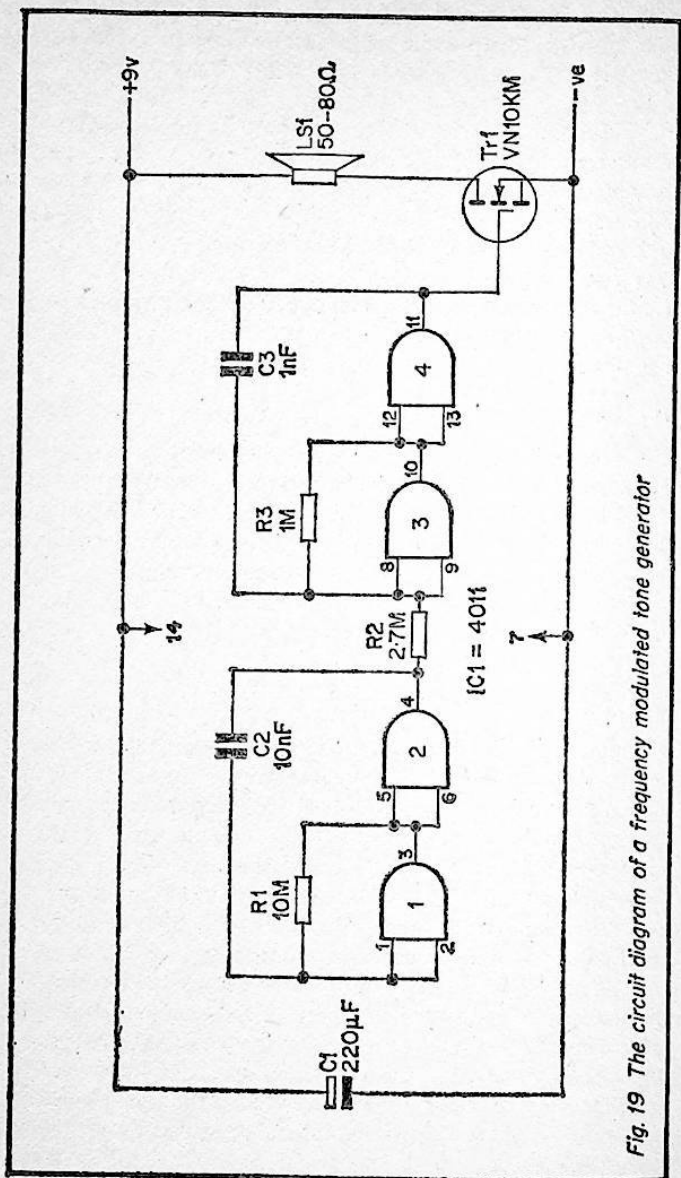


Fig. 19 The circuit diagram of a frequency modulated tone generator

However, R2 loosely couples the output of the modulation astable to the input of the tone generator astable, and the operating frequency of the latter is switched up and down in frequency as the output of the modulation oscillator switches from one logic state to the other. Thus the output of the frequency of the tone generator is actually switched between about 500 Hertz and 1.2 kHz, and it does not operate at its nominal operating frequency at all. This gives a very effective output signal.

As was the case with the previous design, the operating frequencies of the modulation and tone generator astables can be altered to suit individual tastes by changing the values of C2 and C3.

The current consumption of the circuit is about 70mA, and the output power is approximately 300mW.

Components: Frequency Modulator Tone Generator (Figure 19)

Resistors (all 1/3 watt 10%)

R1	10M	R3	1M
R2	2.7M		

Capacitors

C1	220 μ F 16V	C3	1nF plastic foil
C2	10nF plastic foil		

Semiconductors

IC1	4011
Tr1	VN10KM (or VN66AF, VN67AF, etc.)

Miscellaneous

Component board, high impedance (50 to 80 ohm impedance)
loudspeaker, wire, solder, etc.

High Power Version

Like the previous circuit, the circuit of Figure 19 can be modified to give higher output power by changing Tr1 to a VN66AF device fitted with a small finned heatsink, increasing the supply voltage to 12 volts, and using an 8 or 15 ohm impedance loudspeaker. Using an 8 ohm loudspeaker the current consumption is around 600mA and the typical output power is about 3.5 watts. These figures become about 350mA and 2 watts respectively if a 15 ohm impedance loudspeaker is used.

In theory it would be possible to obtain even higher output power by using two VN66AF devices in parallel in the output stage (or a high power VMOS device such as the VN64GA) together with a 4 or 5 ohm impedance loudspeaker and a 12 volt supply. This would give an output power of around 6 watts with an average current consumption of approximately 1.2 amperes (frequency modulated version) or 600mA (pulsed tone circuit). However, in practice it might prove difficult to obtain a 4 or 5 ohm impedance speaker of adequate power rating. This is a point which should be borne in mind with all the tone generator circuits (and the audio circuits of the previous chapter for that matter) since they are quite capable of seriously damaging loudspeakers of inadequate power rating! The speakers should have an RMS power rating at least as high as the specified output power of the circuit concerned, and should preferably have a power rating comfortably in excess of this.

CHAPTER 4

DC CONTROL CIRCUITS

In DC control applications the high operating speed, lack of thermal runaway, and other advantages of VMOS devices over bipolar devices are often of little or no importance. However, the low drive current requirement of VMOS devices does often enable simplified circuits to be used in DC applications when these devices are employed, and they do sometimes offer a superior alternative to bipolar transistors. They are certainly no less viable in such applications than bipolar alternatives.

In this chapter a number of simple DC control circuits such as DC motor speed controllers and a touch switch will be described, and these illustrate the types of circuit in which VMOS devices can be used advantageously.

Touch Switch

The circuit diagram of Figure 20 shows how a simple but very efficient touch switch can be produced using a CMOS IC in conjunction with a VMOS transistor. The circuit is intended to provide on/off switching for small 9 volt DC loads, such as transistor radios, but it will work properly on supply voltages from 6 to 12 volts, and can also be used to control higher current loads as we shall see later.

The circuit is based on two of the four gates in a 4011 CMOS quad 2 input NAND gate IC. The two gates that are used in the circuit each have their two inputs connected together so that they actually provide a simple inverter action. These are then connected in series so that the output of gate 2 assumes the same logic state as the input of gate 1. R1 provides DC positive feedback from the output of the circuit to the input, so that if, for example, the input is taken to the high logic state and then left floating, the output goes high and then holds the input in this state due to the coupling via R1. Thus the circuit tends to latch in whatever state it is left in (provided this is a stable state and not during a transition), and it is actually a simple form of bistable multivibrator.

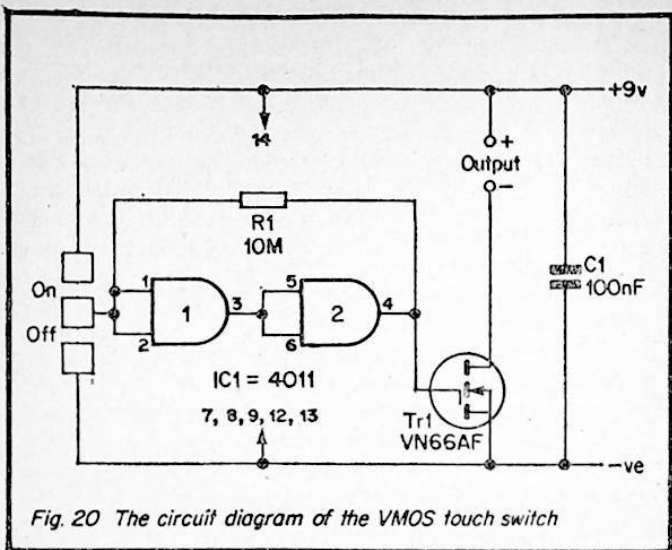


Fig. 20 The circuit diagram of the VMOS touch switch

The bistable can be switched to the desired state by operating the appropriate pair of touch plates. If we assume that the output is initially in the low state so that Tr1 and the load are both switched off, touching the upper two touch plates places a resistance (the resistance through the skin of the operator's finger tip) from the input of the bistable to the positive supply rail. Although this resistance is likely to be in the region of one or two megohms, it is still sufficiently low to take the input of the bistable to more than 70% of the supply potential (which in CMOS circuitry constitutes the high logic state). This occurs due to the very high input impedance of the bistable which is about 0.75 million megohms, and due to the high value of R1. The output of the bistable therefore triggers to the high state, switching on Tr1 and the controlled equipment which is used as its drain load. Of course, the circuit latches in this state when the operator's finger is removed from the touch contacts.

In order to switch the circuit back to the off state again it is merely necessary to touch the lower two touch contacts so that the skin resistance of the operator's finger pulls the input

of the bistable down to the low logic state, latching the circuit in this state.

The circuit is very efficient since the CMOS IC has a negligible current consumption when resting in either logic state, and only consumes a brief pulse of current when making the transition from one logic state to the other. The inputs of the two unused gates (pins 8, 9, 12 and 13 of IC1) are connected to the negative supply rail so that they cannot be operated by stray pick up, and the current consumption of the unused gates is therefore insignificant. Even when switched hard on, Tr1 requires no significant gate current, and so no power is wasted here. When Tr1 is switched off it is only likely to pass a leakage current of about $1 \mu\text{A}$ or less, and no significant power is wasted here either.

A small amount of power will be dissipated in Tr1 when it is switched to the on state, but at a drain current of 100mA it will typically only provide a voltage drop of 200mV, and this is not enough to be of significance in the majority of applications. At higher load currents the voltage dropped across Tr1 does increase proportionately, and would be about 2 volts at 1 ampere for example. It is therefore advisable to only use the circuit in applications where the maximum drain current will be about 300mA or less. If it is used with a supply voltage of only about 6 volts or so, the gate to source voltage of Tr1 when it is in the on state will obviously be reduced, giving Tr1 increased "on" resistance. It is then advisable to keep the load current to no more than about 200mA.

The circuit can be given improved maximum current handling ability by using two or three VN66AF devices wired in parallel in the Tr1 position. This would reduce the effective "on" resistance of the output stage by a factor of two and three respectively, enabling two or three times as much drain current to be passed for a given voltage drop. For even higher load currents Tr1 could be replaced by a high power VMOS device having a very low drain to source "on" resistance.

Components: VMOS Touch Switch (Figure 20)

Resistor (1/3 watt 10%)

R1 10M

Capacitor

C1 100nF plastic foil

Semiconductors

IC1 4011

Tr1 VN66AF

Miscellaneous

Component panel, touch contacts, wire, solder, etc.

Motor Speed Controller

The very simple DC motor speed controller circuit of Figure 21 uses a VMOS transistor as a sort of high power potentiometer (rheostat). The circuit is intended for use with 12 volt DC motors having a maximum current consumption of up to about 1 ampere.

The mains supply is taken to the primary winding of isolation and step-down transformer T1 by way of on/off switch S1. The output of T1 is full wave rectified by the push-pull rectifier circuit comprised of D1 and D2, and the resultant rough DC output is then smoothed to some degree by C1 to give a reasonably steady DC potential. There is in fact a fair amount of ripple on this DC output, but this is of no consequence in this application.

Power is supplied to the load via Tr1, and this is biased by a potential divider circuit which consists of R1, VR1 and R2. With the slider of VR1 at the R2 end of its track, the gate bias voltage fed to Tr1 will not be sufficient to cause this device to conduct significantly, and the motor will not run. Steadily moving the slider of VR1 towards the other end of its track causes a gradually increasing bias to be fed to Tr1, and this results in its drain to source resistance progressively falling. Thus the power fed to the motor increases accordingly, as does the speed of the motor, until Tr1 saturates (at which point the motor is operating at full speed). Thus the speed of the motor can be varied from zero to full speed by means of VR1.

C2 filters out any mains hum or other electrical noise which might otherwise be picked up in the high impedance gate circuit of Tr1, making it impossible to adjust the motor speed down to zero. D3 is a protection diode that suppresses any high reverse voltage spikes that happen to be generated across

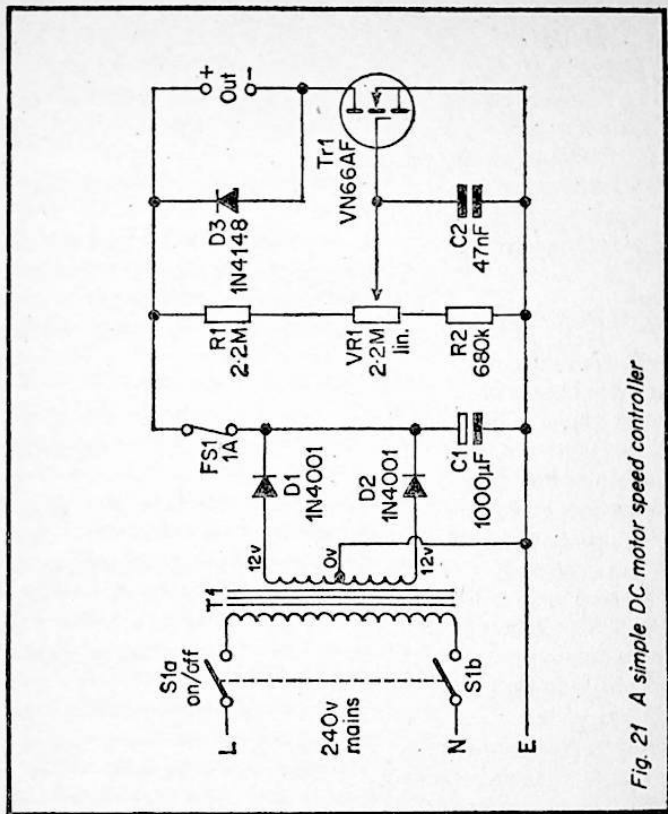


Fig. 21 A simple DC motor speed controller

the highly inductive load provided by the motor.

Components: Simple DC Motor Speed Controller (Figure 21)

Resistors (all 1/3 watt)

R1 2.2M 10%

R2 680k 5%

VR1 2.2M (or 2M) 1in. carbon

Capacitors

C1 1000 μ F 25V

C2 47nF plastic foil

Semiconductors

Tr1 VN66AF

D1 1N4001

D3 1N4148

D2 1N4001

Switch

S1 DP mains switch

Transformer

T1 Standard mains primary, 12V - 0 - 12V @ 1A secondary

Miscellaneous

Component panel, control knobs, case, wire, solder, 1A quick blow fuse and suitable holder, etc.

Pulsed Motor Speed Controller

Although simple motor speed controller designs of the type shown in Figure 21 have the advantage of simplicity, they also have a couple of disadvantages. One of these is simply that there is quite a high level of dissipation in the transistor, especially when the motor is adjusted for about half maximum speed. This is obviously not a major problem though, and merely necessitates the fitting of a reasonably large heatsink to the transistor.

A more serious problem is that the motor has a tendency to stall when this type of controller is adjusted for a low speed. This is due to the fact that the transistor then has a comparatively high resistance, and thus gives the supply a quite high output impedance. If the loading of the motor increases it tries to draw more supply current, but increased supply current results in a higher voltage drop across the transistor and a lower supply voltage across the motor. Therefore the power fed to the motor does not actually change much, and could actually decrease. Thus the motor has a tendency to stall.

There is also an opposite effect where decreased loading on the motor reduces its current drain, causing increased supply voltage and a substantial increase in the speed of the motor.

Much improved motor speed regulation can be attained using a controller that feeds the motor with a pulsed signal. One method of this type, and the one used here, is to have a circuit that gives a fixed output pulse duration, and varied the motor speed by varying the frequency of the pulses. A low frequency gives large intervals between the pulses and a low average power is fed to the motor. A low motor speed is then

obtained. A high frequency results in no significant gaps between the pulses, and virtually a continuous signal is fed to the motor. This gives a high average power in the motor, and it operates at full speed.

The advantage of this system is that while the motor is being pulsed with power it is effectively operating at full power, and can draw a high supply current if the loading on the motor requires it. The motor is therefore operated by a series of powerful pulses which resist any tendency to stall and give high torque even at low speeds.

Figure 22 shows the circuit diagram of a DC motor speed control that uses a pulse method of operation. A suitable DC supply is derived from the mains input by T1, D1, D2 and C1, as in the previous circuit. Again Tr1 is connected in series with the motor, but this time its gate terminal is fed with the output signal from an astable multivibrator circuit. This is based on two of the four gates in a CMOS 4001 device, and these are used in what is almost a standard CMOS astable configuration. It differs from the standard configuration in that there are two timing resistors connected between the output of gate 1 and the junction of R1 and C2. These two resistors are VR1 and R2, and steering diodes D3 and D4 are included in series with these. Due to the presence of these diodes, R2 effectively becomes the timing resistance when the output of the astable is high, and VR1 acts as the timing resistance when the output is low. As R2 has a fixed value the duration of the output pulses is fixed. The period between them is obviously variable by means of VR1, and is virtually zero if it is set for minimum resistance. At maximum resistance the mark space ratio of the output is more than ten to one. VR1 can therefore be adjusted for the required motor speed, with minimum speed occurring at full resistance, and maximum speed being produced at zero resistance.

Like the previous circuit, this controller is intended for use with 12 volt DC motors which draw a supply current of up to about 1 ampere.

Components: Pulsed Motor Speed Controller (Figure 22)

Resistors (both 1/3 watt 5%)

R1 680k

R2 220k

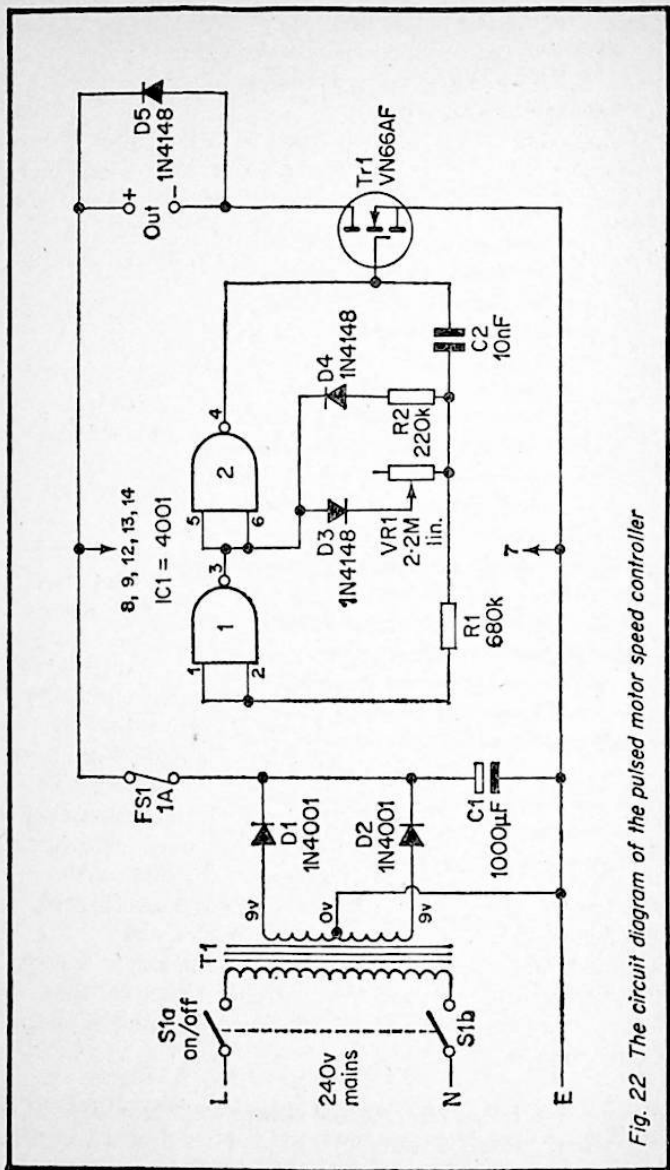


Fig. 22 The circuit diagram of the pulsed motor speed controller

VR1 2.2M (or 2M) 1in. carbon

Capacitor

C1 1000 μ F 25V

C2 10nF

Semiconductors

IC1 4001

Tr1 VN66AF

D1 1N4001

D4 1N4148

D2 1N4001

D5 1N4148

D3 1N4148

Switch

S1 DP mains switch

Transformer

T1 Standard mains primary, 9V - 0 - 9V 2A secondary

Miscellaneous

Component board, control knobs, case, 1A quick blow fuse and holder to suit, wire, solder, etc.

Car Cassette Supply

The circuit of Figure 23 is for a device which enables a cassette recorder or similar item of equipment to be powered from a 12 volt car battery. This is much more economic than using ordinary dry cells to power equipment used in the car, but there is the problem that most cassette recorders require a 6, 7.5 or 9 volt supply. A device such as the one described here is therefore needed in order to drop the 12 volt supply to a suitable level. With the specified values this circuit gives an output potential of 7.5 volts, but this can be changed to approximately 6 or 9 volts by changing D2 to a 6.2 volt or 9.1 volt component respectively. The unit has built-in current limiting circuitry which limits the maximum output current to about 500mA or so.

The circuit uses what is nearly a conventional arrangement, but whereas an emitter follower (or source follower) stage would normally be added at the output of operational amplifier IC1 in order to boost its maximum output current capability to a suitably high level, a common source stage is used here (equivalent to a bipolar common emitter stage). IC1 and Tr3 are used as a simple unity gain buffer stage with 100% negative feedback from the output to the non-inverting input

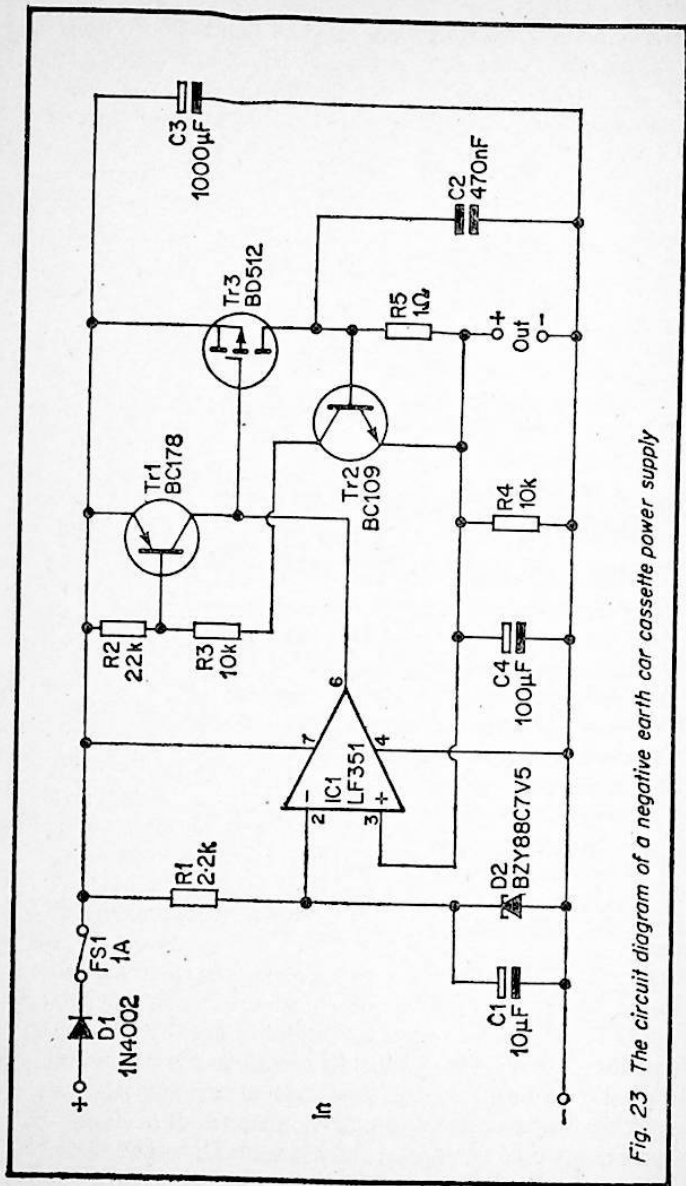


Fig. 23 The circuit diagram of a negative earth car cassette power supply

of IC1. Note that the feedback must be applied to the non-inverting input and not, as would normally be the case, to the inverting input, due to the phase inversion produced by a common source stage. Since there is unity voltage gain from the input to the output of the amplifier, by setting the input at a stable potential of 7.5 volts using a simple zener shunt stabiliser circuit (R1, D2 and C1), the output is set at the same level. Of course, whereas the input signal is at a high impedance, the output is at a very low impedance and can provide the quite high output currents that are needed in this application.

In a conventional current limiting circuit the collector of Tr2 would be connected to the gate of Tr3, and R2, R3 and Tr1 would be unnecessary. This method would not work here, since at high output currents when the voltage developed across R5 was sufficient to switch on Tr2, additional gate biasing for Tr3 would be provided by Tr2, causing Tr3 to conduct more heavily and increase the output current. This is the exact opposite of what is required!

The additional components are therefore required. Now when Tr2 is biased into conduction it switches on Tr1, which then reduces the gate bias on Tr3 and gives the required output current limiting. Due to the very high gain provided by Tr1 and Tr2 there is practically no increase at all in the output current once the current limiting has commenced.

Normally with a current limit resistor value of 1 ohm the maximum output current of a circuit would be about 650mA or so, but in this case the output short circuit current is only about 500mA. This difference occurs because in a normal current limiting circuit Tr2 would be biased hard into conduction under short circuit conditions. In this circuit Tr2 has to pass only the small collector current needed in order to bias Tr1 hard into conduction. Thus the voltage produced across R5 is only about 0.5 volts or so, rather than the more usual figure of around 0.65 volts, and the short circuit output current is correspondingly lower.

The four capacitors are needed in order to filter out noise spikes and to aid the stability of the circuit. R4 is a load resistor for Tr3. D1 protects the circuit in the event of it being connected to the 12 volt supply with the wrong polarity.

The current consumption of the circuit is about 2mA plus whatever output current is drawn. Tr3 should be fitted with a heatsink, and a small, finned, bolt-on type will normally be adequate.

The reason for using a common source output stage in this circuit rather than a source follower should perhaps be explained. As we saw earlier in the chapter describing audio circuits, a source follower output stage produces a substantial voltage drop between the input signal to the gate and the output signal from the source. In this application we do, of course, require a fairly large voltage drop from the input to the output, but the voltage drop through a source follower stage could still be too high, giving an inadequate output potential at high supply currents. In audio circuits this problem can be overcome quite often by the use of the bootstrapping technique, as we have already seen, but this technique can only be used in AC circuits and is not applicable here. It is possible to provide a stepped-up supply for the driver circuitry and thus overcome the problem, but this would be relatively complex, and the use of a common source output stage is probably the most simple way of overcoming the problem.

*Components: Negative Earth Car Cassette Power Supply
(Figure 23)*

Resistors (all 1/3 watt 5% except R5 which is 1 watt 5%)

R1	2.2k	R4	10k
R2	22k	R5	1 ohm
R3	10k		

Capacitors

C1	10 μ F 10V	C3	1000 μ F 25V
C2	470nF plastic foil	C4	100 μ F 16V

Semiconductors

IC1	LF351		
Tr1	BC178	Tr3	BD512
Tr2	BC109		
D1	1N4002	D2	BZY88C7V5 (see text)

Miscellaneous

Component board, 1A quick blow fuse and holder to suit, wire, solder, etc.

Positive Earth Version

A positive earth version of the care cassette power supply is shown in Figure 24. This is basically the same as the negative earth version, and is just rearranged slightly to suit the

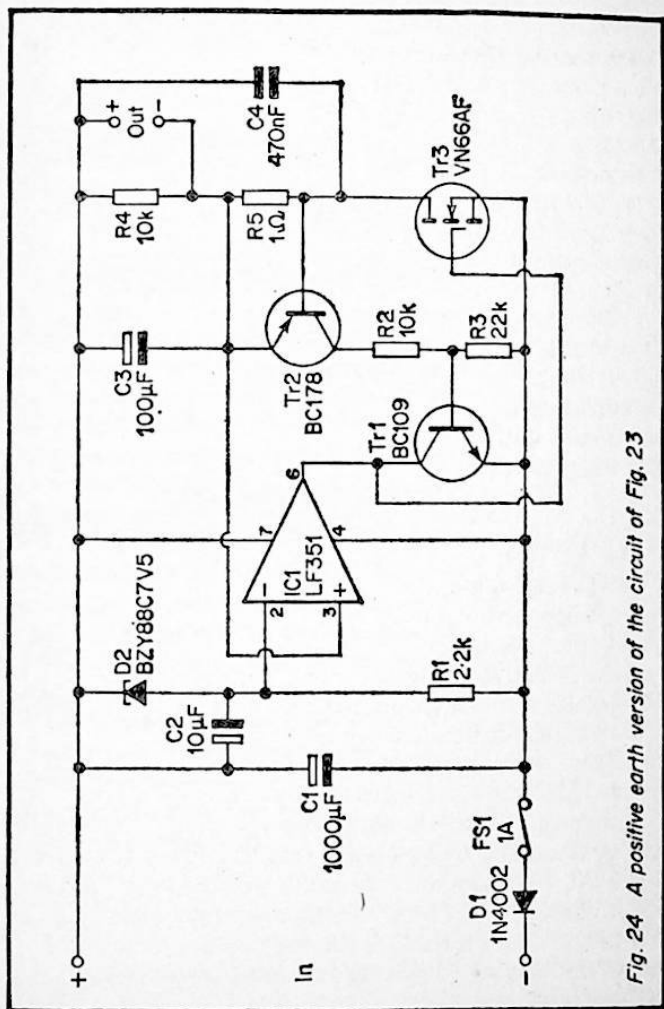


Fig. 24 A positive earth version of the circuit of Fig. 23

change of polarity. For instance, Tr1 has been changed to an NPN device, Tr2 is now a PNP device, and Tr3 is an N channel VMOS device. The circuit is fundamentally identical to the original though.

*Components: Positive Earth Car Cassette Power Supply
(Figure 24)*

Resistors (all 1/3 watt 5% except R5 which is 1 watt 5%)

R1	2.2k	R4	10k
R2	10k	R5	1 ohm
R3	22k		

Capacitors

C1	1000 μ F 25V	C3	100 μ F 16V
C2	10 μ F 10V	C4	470nF plastic foil

Semiconductors

IC1	LF351		
Tr1	BC109	Tr3	VN66AF
Tr2	BC178		
D1	1N4002	D2	BZY88C7V5 (see text)

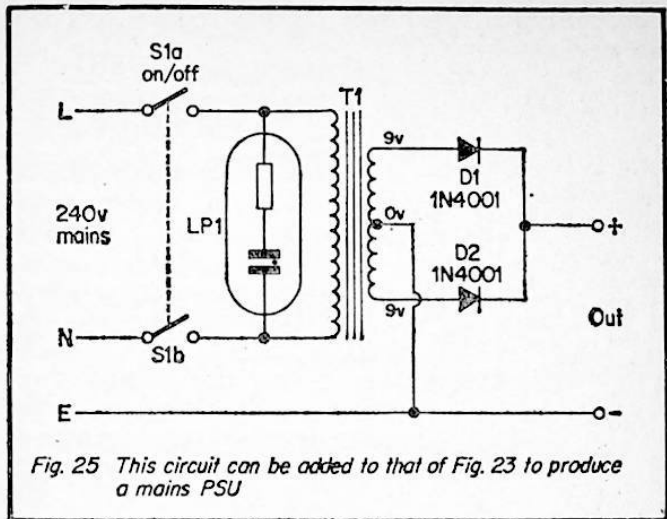
Miscellaneous

Component board, case, 1A quick blow fuse and holder to suit, wire, solder, etc.

Mains Version

By adding a step-down, isolation, and rectifier circuit at the input of either version of the car cassette power supply, a useful battery eliminator is obtained. Suitable additional circuitry for Figure 23 is shown in Figure 25.

The mains supply is coupled to the primary winding of step-down and isolation transformer T1 by way of on/off switch S1. LP1 is a neon on/off indicator, and it is essential that this should be a type which has a built-in series resistor for use on the normal 240 volt UK mains supply. The output of T1 is full wave rectified by a push-pull rectifier circuit which uses D1 and D2. No smoothing capacitor is included in this circuit as C3 in the circuit of Figure 23 will provide adequate smoothing. D1 of Figure 23 is not really needed if the unit is used as a mains power supply, and it can simply be replaced with a shorting wire.



The circuit of Figure 25 can also be used in conjunction with the circuit of Figure 24, but the mains earth lead should then be connected to the positive supply rail instead of the negative one.

Components: Mains Power Supply Unit (Figure 25)

Components as for Figure 23 or Figure 24 plus the following additional components

S1 DP mains switch

LP1 Panel indicator neon having integral series resistor for 240V mains use

T1 Standard mains primary, 9V – 0 – 9V @ 1A secondary

D1 1N4001

D2 1N4001

Control knob, wire, solder etc.

Variable Voltage Supply

The combined circuits of Figure 23 and Figure 25 can easily be modified to perform as a simple variable voltage bench power supply, and Figure 26 details the necessary changes (which only affect the circuit of Figure 23).

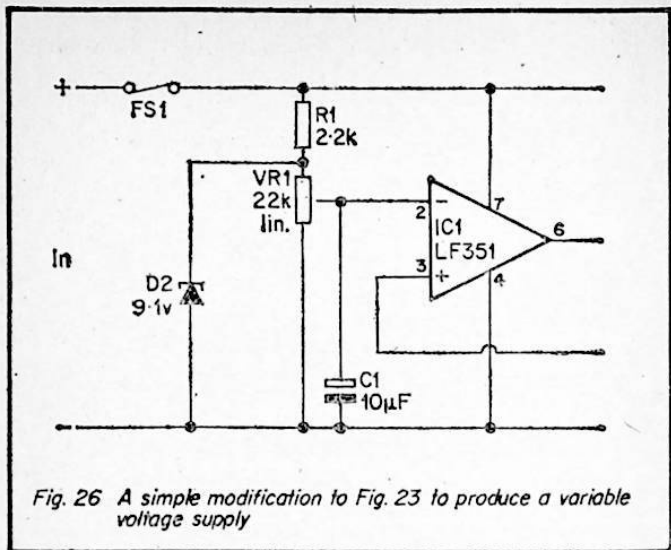


Fig. 26 A simple modification to Fig. 23 to produce a variable voltage supply

All that has been done here is to change D2 to a 9.1 volt component, and to add a potentiometer between the stabiliser circuit and the inverting input of IC1. VR1 can therefore be adjusted to give a stable potential of anything between zero and 9.1 volts at the inverting input of IC1, and the output is therefore adjustable between these same limits.

Of course, the circuit of Figure 24 can be modified in a similar manner if desired. When these circuits are used as a bench power supply it is quite in order to take the mains earth to an output socket, and to connect neither output terminal direct to the mains earth. Then the desired output terminal can be earthed, or neither need be earthed, as circumstances dictate.

In any event, in the interests of safety, any exposed metal-work other than the output terminals should be connected reliably to the mains earth.

It is very helpful if VR1 is given a scale calibrated in terms of output voltage, or a voltmeter could of course be used at the output of the unit to indicate the output potential.

Components: Variable Voltage Power Supply (Figure 26)
As for fixed voltage supply plus 22k lin. carbon potentiometer (VR1)

Automatic Parking Light

This simple circuit will automatically switch on a parking light at dusk, and switch it off again at dawn. The light level at which the switching action occurs is adjustable over a wide range. The complete circuit diagram of the unit is shown in Figure 27.

The circuit is based on an operational amplifier which is connected so that it operates as a sort of Schmitt Trigger. A VMOS device is used to interface the relatively high output impedance of the operational amplifier to the low impedance bulb load.

R1 is used to supply some fraction of the supply voltage to the inverting input of IC1, and PCC1 plus R2 form a potential divider circuit which supplies a fraction of the supply

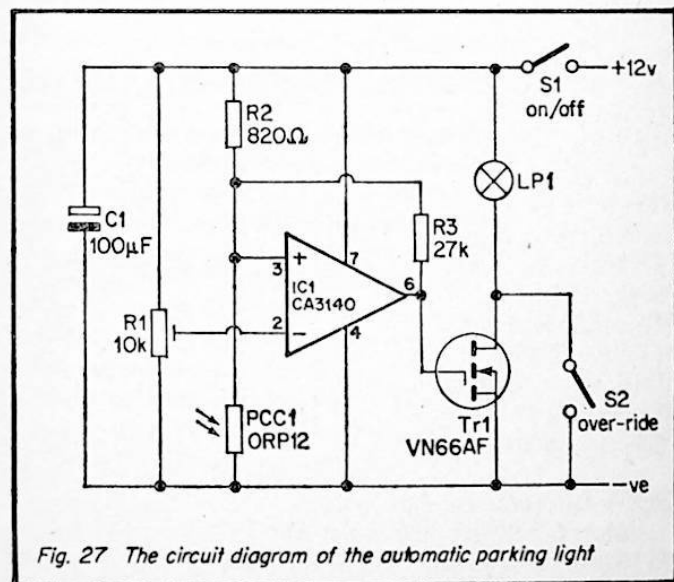


Fig. 27 The circuit diagram of the automatic parking light

potential to the non-inverting input. The voltage provided by the latter depends upon the light intensity received by PCC1 which is a cadmium sulphide photo-resistor, and has a resistance that falls with increasing light level. Thus the voltage fed to the non-inverting input of IC1 is only a fraction of a volt under very high levels of illumination, and is virtually equal to the full supply potential under low light levels.

R1 is adjusted so that under the kind of light levels present at dusk and dawn the two input voltages to IC1 are roughly balanced. Thus, during normal daylight the voltage fed to the non-inverting input is lower than that supplied to the inverting input, causing the output of IC1 to go low. Tr1 is therefore switched off, as is the bulb of the parking light which forms its drain load. As the light level falls as dusk approaches, the voltage fed to the non-inverting input gradually reduces until it is lower than the reference level at the inverting input. The output of IC1 then goes high, switching on both Tr1 and LP1. At dawn the voltage fed to the non-inverting input rises once again until it is more than the reference level at the inverting input, resulting in the output of IC1 going low, and both Tr1 and LP1 being switched off. The required action is therefore obtained, with the light being switched on at dusk and off again at dawn.

R3 provides a small amount of positive feedback that ensures that IC1 switches cleanly and rapidly from one output state to the other, without instability and erratic operation. S2 enables the unit to be over-ridden so that LP1 can be switched on even if the light level is above the switch-on threshold level. S1 is an ordinary on/off switch.

Since Tr1 is either switched hard on or cut off it does not have to dissipate much power and should not require any heatsinking. Note that PCC1 should not be mounted where it will pick-up a significant amount of light from LP1, or under dark operating conditions the circuit may start to oscillate due to the feedback between these two components!

Components: Automatic Parking Light (Figure 27)

Resistors (all 1/3 watt 5% except R1)

R1	10k 0.1 watt preset	R3	27k
R2	820 ohms		

PCC1 ORP12 CDS cell

Capacitor

C1 100 μ F 25V

Semiconductors

IC1 CA3140T or CA3140E

Tr1 VN66AF

Switches

S1 SPST toggle

S2 SPST toggle

LP1 12 volt parking lamp bulb

Miscellaneous

Case, component board, wire, solder, etc.

Ultra Simple Timer

The high input impedance of VMOS devices together with their enhancement mode of operation makes them ideal for timer circuits. The circuit of Figure 28 shows how a VMOS transistor can be used as the basis of a very simple switch on delay timer, where power is not supplied to a relay until some predetermined period after the unit has been switched on. Circuits of this type can be used in burglar alarms and many other applications.

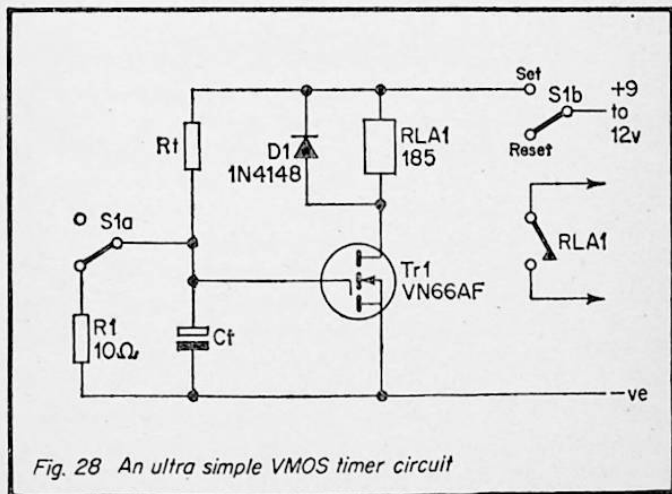


Fig. 28 An ultra simple VMOS timer circuit

When power is first applied to the circuit Tr1 will be cut off, and will not supply any significant current to the relay which acts as its drain load. Ct immediately starts to charge via Rt, and eventually Tr1 will begin to conduct. When the drain current reaches a suitably high level the relay is activated. When the circuit is reset, S1a discharges Ct through current limiting resistor R1 so that the unit is ready to perform another timing run when S1 is returned to the "set" position again. R1 is needed to eliminate sparking at the contacts of S1a due to an excessive discharge current (which would reduce the operating life of S1 substantially).

The delay time is approximately $0.3CR$ seconds with C in micro farads and R in megohms, but this is only a very approximate figure as in practice such things as the relay threshold current, threshold voltage of Tr1, and the precise supply voltage all affect the time delay. This circuit is therefore only suitable for applications where highly predictable output times are required. Rt can be made many tens of megohms in value if necessary, and so quite long delay times are attainable. If a high value is used for Rt, and Ct is an electrolytic component, it is advisable to use a tantalum bead component for the latter. An ordinary electrolytic may have a leakage current that is too high for this application, greatly extending the timing period or even preventing the relay from switching on at all.

The circuit can easily be modified to give an alternative form of operation where switching on causes the relay to be energised for some predetermined length of time. The necessary modification is detailed in Figure 29. Note that the original circuit gave a relatively short timing period due to the fact that Ct only had to charge to a small fraction of the supply voltage in order to switch on Tr1. In the modified circuit Ct has to charge to not far short of the full supply potential before Tr1 starts to switch off and the timing period ends. This gives a switch-on time of very roughly $2.5CR$. It also makes it even more important for Ct to be a high quality low leakage component.

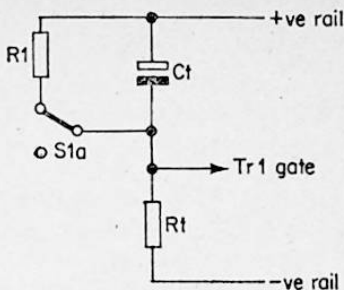


Fig. 29 The circuit of Fig. 28 can be modified to switch on the relay for a predetermined length of time, as shown here

Components: Ultra Simple VMOS Timer (Figure 28)

Resistors

R1 10 ohms 1/3 watt 5%

Rt see text

Capacitor

Ct see text

Semiconductors

Tr1 VN66AF

D1 1N4148

Switch

S1 DPDT toggle

Relay

RLA1 6/12 volt coil having a resistance of 185 ohms or more, and make contacts of adequate ratings for the intended load

Miscellaneous

Case, component panel, wire, solder, etc.

Radio Timer

The timer circuit of Figure 30 is primarily intended for use with a 9 volt battery operated radio that is used as a bedside set. The timer can be used to automatically switch off the set after a period of 5, 10, 15 or 20 minutes, giving the "sleep"

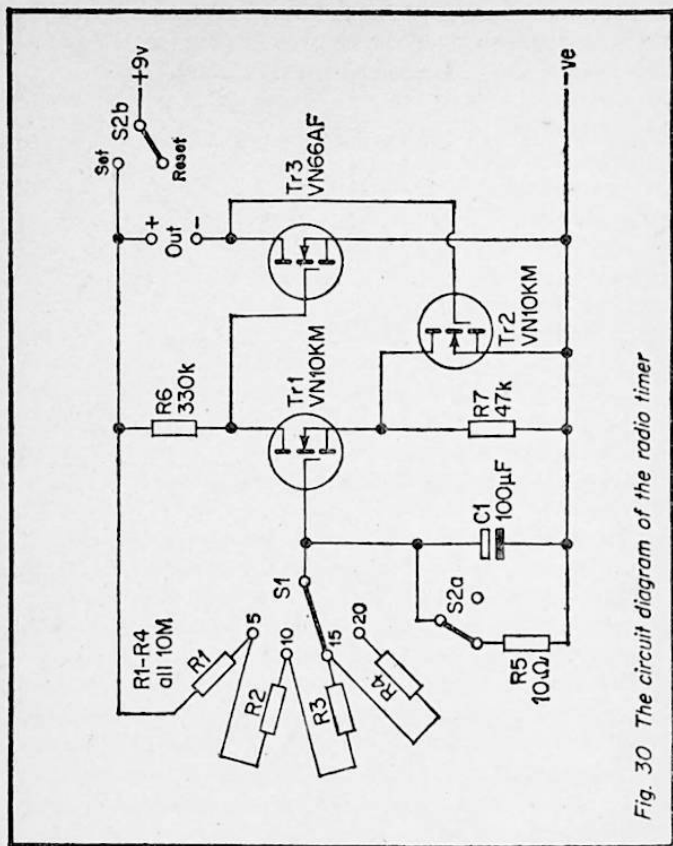


Fig. 30 The circuit diagram of the radio timer

or "snooze" facility that is built into many clock radios and some other sets. As in the previous circuit, the timing periods provided by the unit are not highly accurate and reliable, but this is of no consequence in a non-critical application such as the present one.

At switch-on C1 is uncharged and Tr1 is therefore cut off. R6 biases Tr3 hard into conduction and power is supplied to the radio which forms the drain load of Tr3. The charge on C1 gradually increases until Tr1 starts to conduct. The voltage across R6 then starts to rise, and the voltage across R7 starts

to rise as well. Eventually Tr1 will conduct hard enough to produce a large proportion of the supply voltage across R6, and Tr3 starts to switch off. Its drain potential then rises, and switches on Tr2. This pulls the source terminal of Tr1 lower in potential, increasing the gate to source voltage of this device and causing it to conduct more heavily. This results in Tr3 conducting less heavily, which in turn biases Tr2 harder and further increases the bias to Tr1. This regenerative action occurs very rapidly, and results in Tr1 and Tr2 being switched hard on, and Tr3 being cut off. Power is thus removed from the radio. This regenerative feedback is needed in order to ensure that the radio is switched off abruptly rather than being slowly faded out. Of course, if a slow fade out is preferred, then it is merely necessary to omit Tr2.

It is obviously essential that the circuit should have a low current consumption so that the life of the battery is not greatly reduced (bearing in mind that when the set switches off the timer will still be connected to the battery, and will not be disconnected until the morning). In the off state the only significant current drawn by the set is that which flows through R6, Tr1 and R7. Due to the fairly high combined resistance of these a current flow of only about 23 μ A ensues, and this is far too small to have any large effect on the lifespan of the battery.

The charge time of C1 depends upon the number of timing resistors switched into circuit using S1. A timing period of roughly five minutes is obtained with a single resistor, and consequently times of about 10, 15 and 20 minutes with two, three and four resistors respectively.

When the unit is switched off, S2a discharges C1 through current limiting resistor R5 so that the unit is ready to start another timing run from the beginning when S2 is returned to the "set" position again.

Components: Radio Timer (Figure 30)

Resistors (all 1/3 watt 5%)

R1 to R4	10M (4 off)	R6	330k
R5	10 ohms	R7	47k

Capacitor

C1 100 μ F 10V tantalum bead

Semiconductors

Tr1 VN10KM (or VN66AF, VN67AF, etc.)

Tr2 VN10KM (or VN66AF, VN67AF, etc.)

Tr3 VN66AF

Switch

S1 4-way, 3-pole rotary (only one pole used)

S2 DPDT toggle

Miscellaneous

Case, component panel, control knob, battery and connectors, wire, solder, etc.

Simple Burglar Alarm

A VMOS device makes a good basis for a burglar alarm circuit since its low level of leakage current and insignificant gate drive current requirement make it easy to produce a circuit that has an extremely low stand-by current consumption. This makes battery operation of the unit an economically sound prospect.

The circuit diagram of a basic burglar alarm based on a VMOS transistor is shown in Figure 31. This can be used with normally open (NO) contacts, normally closed (NC) contacts, or both. These are, of course, micro or reed switches fitted to doors or windows, or possibly switch mats. If no normally open contacts are used, both these and R2 are simply omitted from the circuit. If the normally closed contacts are not required, these are left out and R1 is wired direct between Tr2's gate and the negative supply rail.

Operation of the circuit is very straight forward. Under quiescent conditions R1 and R4 provide a bias voltage to Tr2 that is far too small to bring the device into a state of conduction. Therefore no significant power is supplied to the relay which forms the drain load for Tr2. If one of the NO contacts is activated, a potential divider action across R2 and R1 takes the gate of Tr2 several volts positive, biasing the device into conduction and activating the relay. A pair of normally open relay contacts then close and switch on the alarm generator circuit.

Tr1 was originally cut off, since there was no significant voltage across the relay and thus no significant base current fed

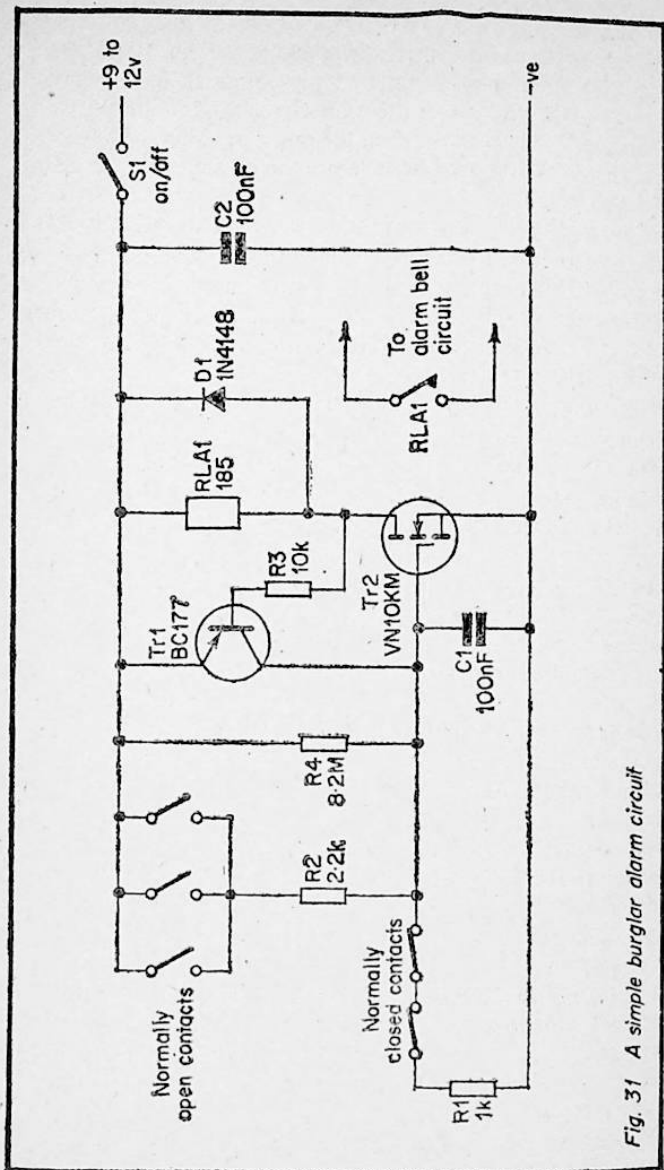


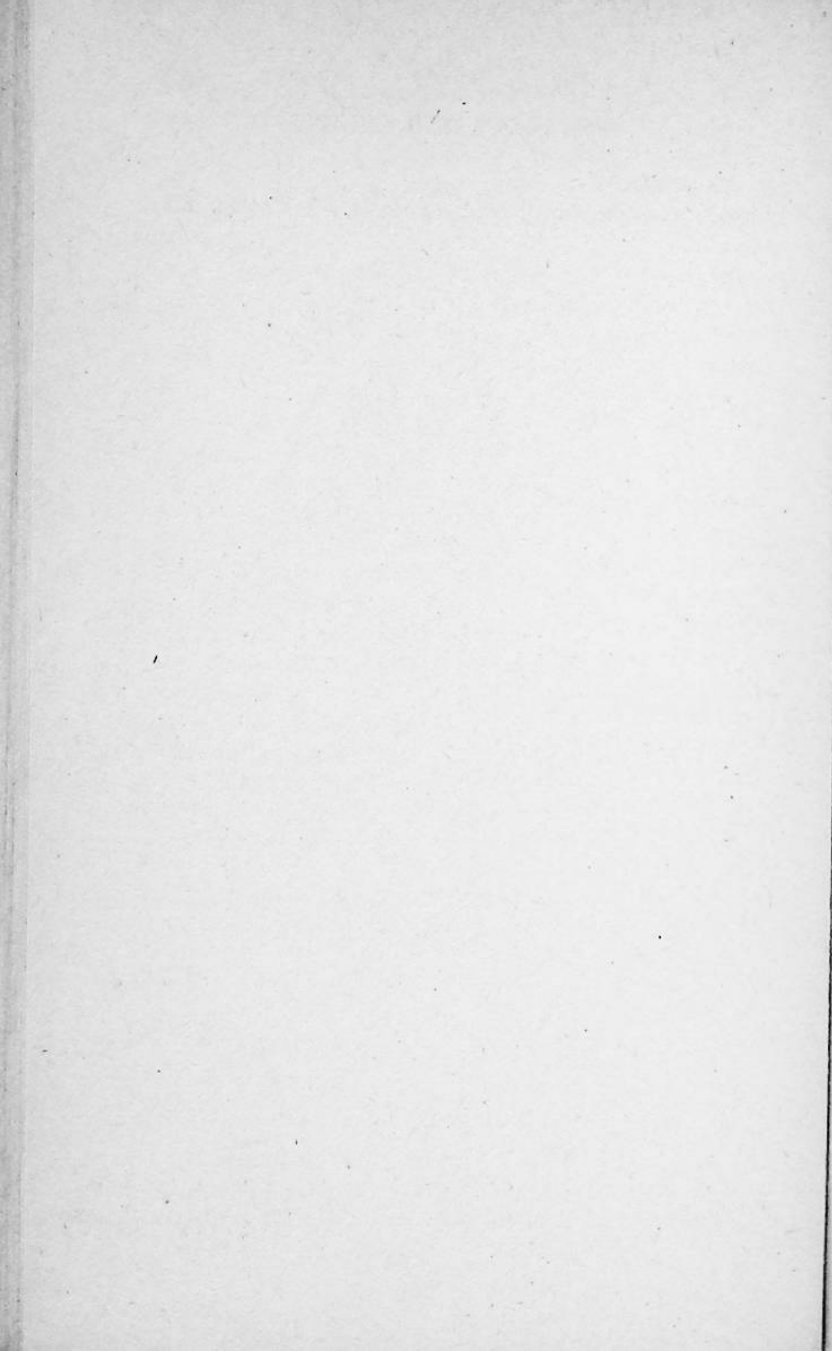
Fig. 31 A simple burglar alarm circuit

Relay

RLA1 6/12 volt coil having a resistance of 185 ohms or more,
and at least one make contact of adequate rating for
alarm generator

Miscellaneous

Strong case, component panel, contacts, wires, solder, etc.



CHAPTER 5

SIGNAL CONTROL CIRCUITS

Like JFETs and ordinary MOSFETs, VMOS devices can be employed as voltage controlled resistors. Being enhancement mode devices, VMOS transistors are often more convenient to use than JFET devices (which are depletion mode devices). The MOSFETs that are generally available are also depletion mode devices, incidentally. Another factor which makes them easy to use is their lower minimum resistance when compared to normal JFET and MOSFET transistors. The latter usually exhibit a minimum resistance of a few hundred ohms — about one hundred times more than that of a typical VMOS device. On the other hand, a JFET or MOSFET device normally has an off resistance of hundreds of megohms or more. Most VMOS devices produce a resistance of many megohms when switched off, but are significantly inferior to JFETs and MOSFETs in this respect. It is therefore advisable to use VMOS transistors in low or medium impedance circuits rather than in high impedance ones when they are utilized as voltage controlled resistors.

As with any FET device, VMOS transistors do not provide a true resistance, in that there is not a strictly linear relationship between applied voltage and resultant current flow. This means that in analogue circuits a certain amount of distortion will inevitably be introduced by the VMOS device. Normally the level of distortion will only be quite low though (only a fraction of 1%). Under certain circumstances a higher distortion level can result, and this usually only occurs if the input signal voltage is comparatively high, and (or) the transistor is biased close to being cut off.

The reason for this will become apparent if one refers back to Figure 3 in Chapter 1. It will be seen from this that the device behaves virtually as an ordinary resistor until a certain current is reached. The current through the component then rises less slowly as the drain to source voltage is increased, until eventually the point is reached where increased voltage has virtually no effect on the level of current flow. In order to obtain

minimum distortion it is obviously necessary to keep the applied voltage below the point at which the transistor begins to saturate. This is really only a problem close to the cut off point where the saturation current and voltage both become quite small.

DC Volume Control

The circuit diagram of Figure 32 is for a DC volume control, and this also illustrates the basic characteristics of a VMOS device when used as a voltage controlled resistor. The advantage of a volume control of this type over a simple potentiometer circuit is that in this circuit the volume control potentiometer only handles a DC signal and there are no problems if this control is located away from the main circuit. Any mains hum or other noise picked up in the cables will be filtered out by C1. It is therefore not even necessary to use a screened connecting cable. Obviously the situation would be very different if the potentiometer was to be used directly as the volume control. A filter capacitor could not be used as it would remove both the controlled signal as well as the hum and

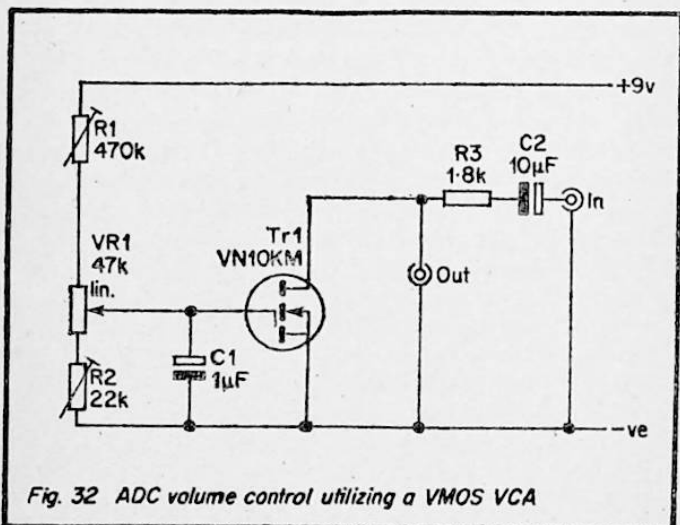


Fig. 32 ADC volume control utilizing a VMOS VCA

noise! It is thus necessary to use screened cable, and due to the capacitance in the cable and the fact that it is less than perfect there are restrictions on the length of cable that can be used in practice.

R3 forms an attenuator in conjunction with the drain to source resistance of Tr1. R1 and R2 are adjusted by trial and error so that Tr1 just becomes cut off when the slider of VR1 is moved to the bottom of its track, and is just biased hard into conduction with VR1's slider moved to the top end of its track. This gives minimum volume with VR1's slider at the top of its track, with the attenuation level through the circuit approaching 60dB. With VR1's slider at the bottom end of the track and Tr1 cut off there is little voltage drop through R3. In fact the voltage drop here will depend almost entirely on the input impedance of the circuit to which the output of the unit is fed.

With this arrangement the voltage across Tr1 is at a maximum when the device is cut off, or close to the cut off point. Although one might expect this to cause high distortion at high volume settings, this is not in fact the case. It must be borne in mind that when Tr1 is close to being cut off its "resistance" is very high in comparison to that of R3, and any variations in the "resistance" due to non-linearity have little effect.

The circuit has an extremely low current consumption of only about 20 μ A.

Components: DC Volume Control (Figure 32)

Resistors

R1	470k 0.1 watt preset	R3	1.8k 1/3 watt 5%
R2	22k 0.1 watt preset		
VR1	47k lin. carbon		

Capacitors

C1	1 μ F 10V	C2	10 μ F 10V
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Semiconductors

Tr1	VN10KM (or VN66AF, VN67AF, etc.)
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Miscellaneous

Component panel, wire, solder, etc.

Tremolo Unit

The tremolo effect is one of the best known musical effects, and merely consists of amplitude modulating the processed signal at a low frequency. In other words the volume of the signal is automatically varied up and down several times per second. Figure 33 shows the circuit diagram of a simple tremolo unit based on a VMOS voltage controlled attenuator (VCA).

Tr1 and R7 form the VCA, and the output from this is fed to a simple common emitter stage based on Tr2. This provides a small amount of voltage gain, and also ensures that the VCA operates into a fairly high impedance so that large losses through R7 when Tr1 is switched off are avoided. R9 introduces negative feedback to Tr2 which gives the required boost in input impedance and prevents Tr2 from having an excessive voltage gain.

IC1 is an operational amplifier which is used in a well known relaxation oscillator configuration. This provides a squarewave output at the output of IC1, and a roughly triangular waveform across C2. A squarewave signal does not make a good modulating signal in this application as it would result in the volume of the processed signal being switched between two levels, whereas a smooth variation in gain is necessary in order to give a good effect. The waveform across C7 is much more suitable, and gives good results in practice.

This signal is at a fairly high impedance, but this does not present any problems due to the high input impedance of Tr1. The signal varies between about $1/3V+$ and $2/3 V+$, and this is a little too high to directly drive Tr1, since it would result in the latter always being switched on fairly hard. R5 and R6 are therefore used to attenuate the signal to a suitable degree. In practice R6 is merely adjusted by trial and error for the best effect.

VR1 controls the tremolo frequency, and the range obtained is from about 0.5 Hertz to nearly 10 Hertz, although this is subject to the wide tolerances of the components used in the modulation oscillator. C3 is the compensation capacitor for IC1, and S1 is a straight forward on/off switch. The circuit has a current consumption of only about 1.5mA.

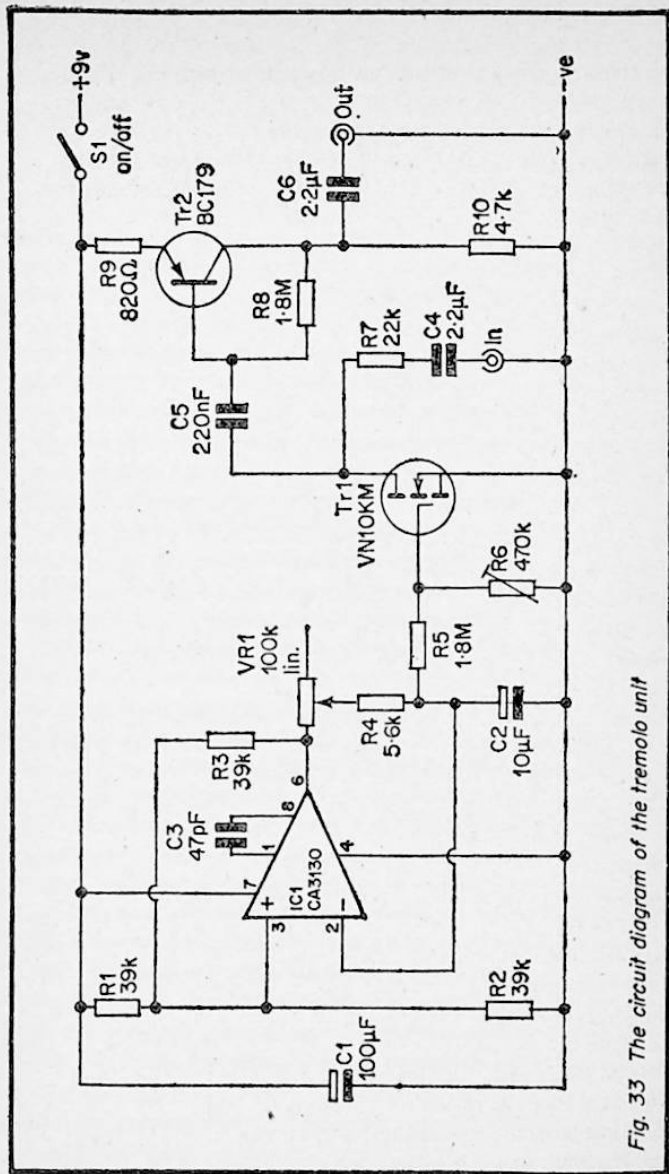


Fig. 33 The circuit diagram of the tremolo unit

The unit is primarily intended for use with a low level signal source such as an electric guitar, but it will work reasonably well provided the input level does not exceed about 500mV RMS.

Components: Tremolo Unit (Figure 33)

Resistors (all 1/3 watt 5% (10% over 1M) except for R6)

R1	39k	R6	470k 0.1 watt preset
R2	39k	R7	22k
R3	39k	R8	1.8M
R4	5.6k	R9	820 ohms
R5	1.8M	R10	4.7k
VR1	100k lin. carbon		

Capacitors

C1	100 μ F 10V	C4	2.2 μ F plastic foil
C2	10 μ F 10V tantalum bead	C5	220nF plastic foil
C3	47pf ceramic plate	C6	2.2 μ F plastic foil

Semiconductors

IC1	CA3130T or CA3130E		
Tr1	VN10KM (or VN66AF, VN67AF, etc.)	Tr2	BC179

Switch

S1	SPST toggle
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Miscellaneous

Case, component panel, 9 volt battery and connector to suit, sockets, wire, solder, etc.

Noise Gate

As many readers may be unfamiliar with noise gates, it should perhaps be explained that this is a piece of equipment which only passes signals that exceed a certain threshold level. Signals are either fully suppressed (giving an effect similar to the squelch circuits fitted to some FM tuners and VHF communications equipment, and circuits of this type are often called "audio squelch" units), or attenuated by about 20dB, if they are below the threshold level. The idea of a circuit of this type is to reduce or eliminate noise on signals which have a poor signal to noise ratio. It is only the noise during gaps in the signal that is effected, and there is no true improvement in the signal to noise ratio. However, a high continuous noise level,

whether electrical in origin, or perhaps something like crowd noise over a public address system, can make signals hard and tiring to listen to for any length of time. By cutting out the noise during pauses in the signal (when the noise would be most obtrusive and noticeable) a noise gate can make a large apparent increase in the signal to noise ratio of a signal.

Although totally suppressing signals below the threshold level might seem to be a good idea, giving the optimum apparent increase in performance, it does have one slight drawback. This is merely that it is possible that the main signal may under certain circumstances just fail to exceed the threshold level, and will then be suppressed. Reducing low level signals by about 20dB is sufficient to render the background noise level unobtrusive, but will leave the main signal audible if it fails to operate the circuit. The circuit described here (and shown in Figure 34) can be used in either mode of operation.

The input signal is fed to the output via R6 and DC blocking capacitors C4 and C5. If Tr3 is switched off, losses through R6 will be minimal and very little attenuation will be provided by the circuit. If, on the other hand, Tr3 is biased hard into conduction, the series resistance of Tr3 and R5 gives losses of about 20dB or so through R6 (if required, R5 can be replaced with a link wire and the attenuation will then be around 60dB, almost totally suppressing the input signal).

Some of the input signal is taken by way of R3 and C6 to a high gain common emitter stage using Tr1. The output from this is coupled by C2 to a rectifier and smoothing circuit which consists of D1, D2 and C3. Normally Tr2 is cut off, and R4 biases Tr3 hard into conduction, the input signal being attenuated in consequence. If the amplitude of the input signal is sufficiently large, the positive bias produced across C3 by the rectified signal will be sufficient to switch Tr2 on, and thus switch off Tr3. The input signal is then allowed to pass virtually unhindered.

This gives the required noise gate action, and R3 is adjusted so that the background noise is suppressed, but the main signal is allowed to pass. The setting of this component should not be very critical unless the difference in the dynamic levels of the main and noise signals is relatively low. Very careful adjustment would then be required in order to give reliable

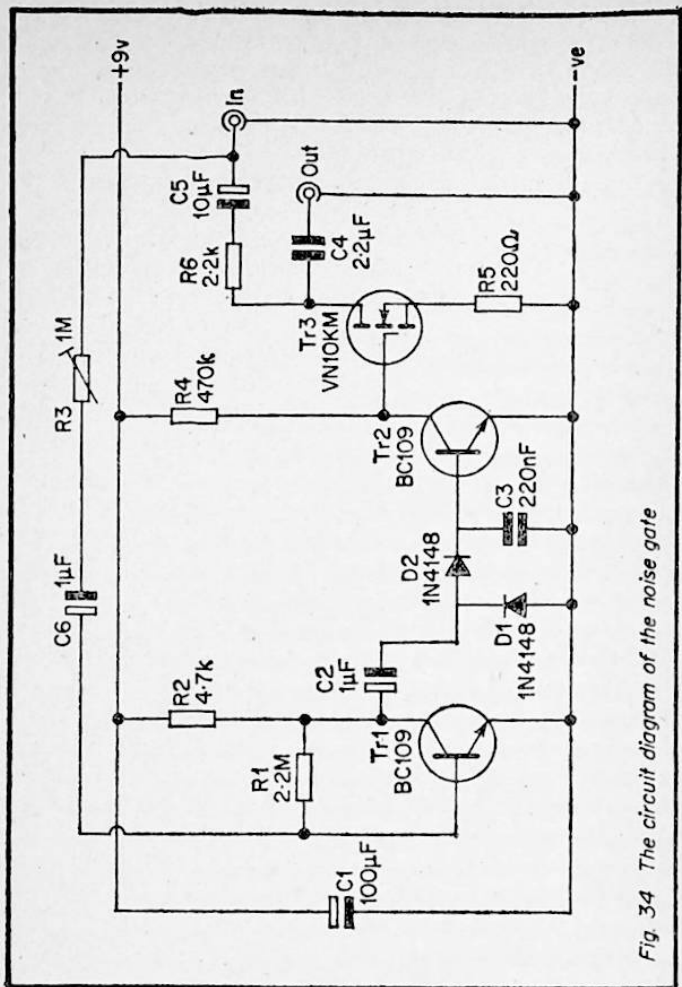


Fig. 34 The circuit diagram of the noise gate

performance. The threshold level of the circuit can be adjusted over very wide limits by means of R3, and the unit is therefore unfussy about the nominal input signal level.

The unit has a current consumption of only about 1mA.

Components Noise Gate (Figure 34)

Resistors (all 1/3 watt 5% (10% over 1M) except R3)

R1	2.2M	R4	470k
R2	4.7k	R5	220 ohms
R3	1M 0.1 watt preset	R6	2.2k

Capacitors

C1	100 μ F 10V	C4	2.2 μ F plastic foil
C2	1 μ F 10V	C5	10 μ F 10V
C3	220nF plastic foil	C6	1 μ F 10V

Semiconductors

Tr1	BC109	Tr3	VN10KM (or VN66AF, VN67AF, etc.)
Tr2	BC109		
D1	1N4148	D2	1N4148

Miscellaneous

Component panel, wire, solder, etc.

Audio Compressor

An audio compressor could be regarded as a form of automatic volume control which comes into operation if the input signal exceeds a certain level, and the circuit gain is then adjusted to prevent the output from more than marginally exceeding the threshold level. Units of this type are used in tape recording to prevent overloading during recording, in transmitting equipment to avoid over-modulation, in music to give the sustain effect when used with an electric guitar, and in many other applications. The unit described here (see Figure 35) has characteristics which make it suitable for use as a sustain unit.

The VCA has R4 as one element and the parallel resistance of Tr2 and R5 as the other. The input signal is coupled to the VCA via a common emitter amplifier which uses Tr1. This provides the unit with a reasonably high input impedance (about 47k) and provides a small amount of voltage amplification as well. Some of the output of the VCA is coupled to the output socket by C4, and the rest is coupled to a high gain common emitter amplifier by R6 and R7.

The output from this amplifier is rectified and smoothed by D1, D2 and C6 to produce a DC control signal for the VCA. Under low input signal conditions the positive bias produced across R7 will not be large enough to switch on Tr2, and there

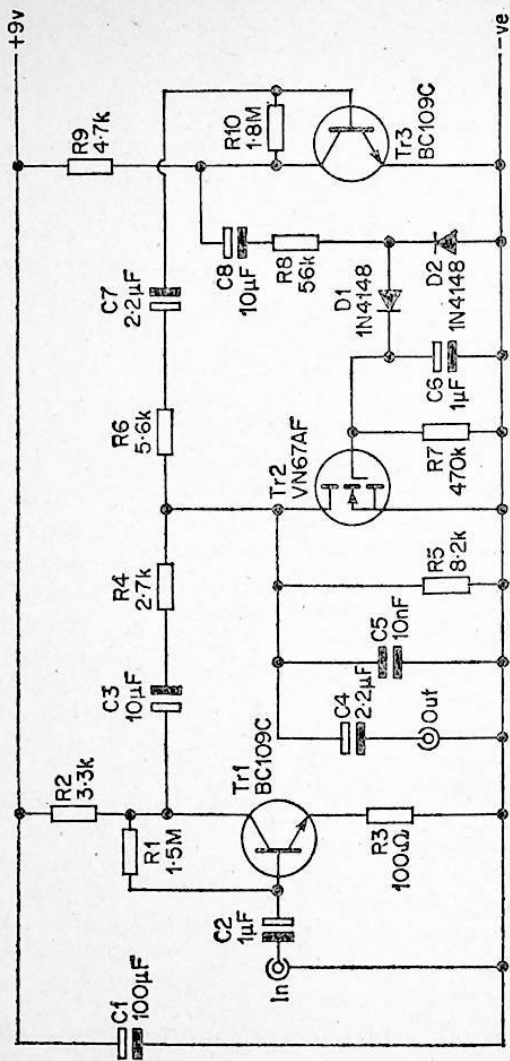


Fig. 35 The circuit diagram of a VMOS audio compressor suitable for use as a sustain unit

will consequently be little attenuation through the VCA. At higher signal levels there will be sufficient bias to switch on Tr2, thus reducing the gain of the circuit. The higher the input signal is taken above this threshold level, the harder Tr1 conducts. This gives the required effect of holding the output level just above the threshold level, and the circuit is actually very effective in this respect due to the high gain of Tr2.

Measurements made on the prototype showed the unit to have a voltage gain of about 18dB (twelve times) for output signals up to about 15mV RMS. Increasing the input to several hundred mV RMS (an increase of over 40dB) caused the output to increase to only about 20mV RMS (an increase of only about 2.5dB)!

R5 is included across Tr2 so that as the latter begins to switch on or off, and is comparatively non-linear, most of the shunt resistance of the VCA is provided by R5, and any non-linearity in Tr2 has relatively little effect. C5 gives a small amount of treble cut at low signal levels, reducing background noise. At higher signal levels where Tr2 exhibits a much lower resistance C5 has very little effect, but at higher signal levels the background noise becomes masked and is not noticeable. Due to the increased gain provided by the unit under low input signal conditions there is inevitably some reduction in the signal to noise ratio of the entire setup, but the noise level should not be very high, and C5 can be omitted if it is considered to be superfluous.

Components: Audio Compressor (Figure 35)

Resistors (all 1/3 watt 5% (10% over 1M))

R1	1.5M	R6	5.6k
R2	3.3k	R7	470k
R3	100 ohms	R8	56k
R4	2.7k	R9	4.7k
R5	8.2k	R10	1.8M

Capacitors

C1	100 μ F 10V	C5	10nF plastic foil
C2	1 μ F 10V	C6	1 μ F 10V
C3	10 μ F 10V	C7	2.2 μ F 10V
C4	2.2 μ F 10V	C8	10 μ F 10V

Semiconductors

Tr1 BC109C

Tr3 BC109C

Tr2 VN67AF or VN66AF

D1 1N4148

D2 1N4148

Miscellaneous

Component panel, wire, solder, etc.

Computer Voice

The circuit shown in Figure 36 can be used to produce computer type voice effects for amateur dramatics productions, etc. Circuits of this type are in many ways similar to a tremolo circuit, but the modulating signal is virtually a squarewave. In this case the modulation signal is provided by a straight forward astable multivibrator which is built around Tr1 and Tr2. A mark space ratio of slightly less than one to one seems to give optimum results, and C4 and C5 are not therefore of equal value, as would be the case for a "text book" astable. Slightly improved results are obtained by slowing up the risetime of the modulation signal slightly, and this is achieved by using a simple R - C low pass filter (R5 and C3) between the output of the astable and the gate of Tr3.

Tr3 is used in a simple VCA which also utilizes R6 and R7. C1 and C2 are merely DC blocking capacitors, and C6 is a supply decoupling capacitor.

The effect of the circuit is to subject the input signals to two levels of attenuation, the higher one being about 26dB more than the lower one. The modulation oscillator switches the circuit between these two level at a rate of several Hertz, and this "breaks up" the signal to give the appropriate effect. Although this circuit uses the most simple and basic method of electronically producing a computer type voice effect, it works quite well in practice. The input signal level can be anything up to a few volts peak to peak without any problems with overloading occurring.

On/off switching is provided by S1. The circuit has a current consumption of about 1mA.

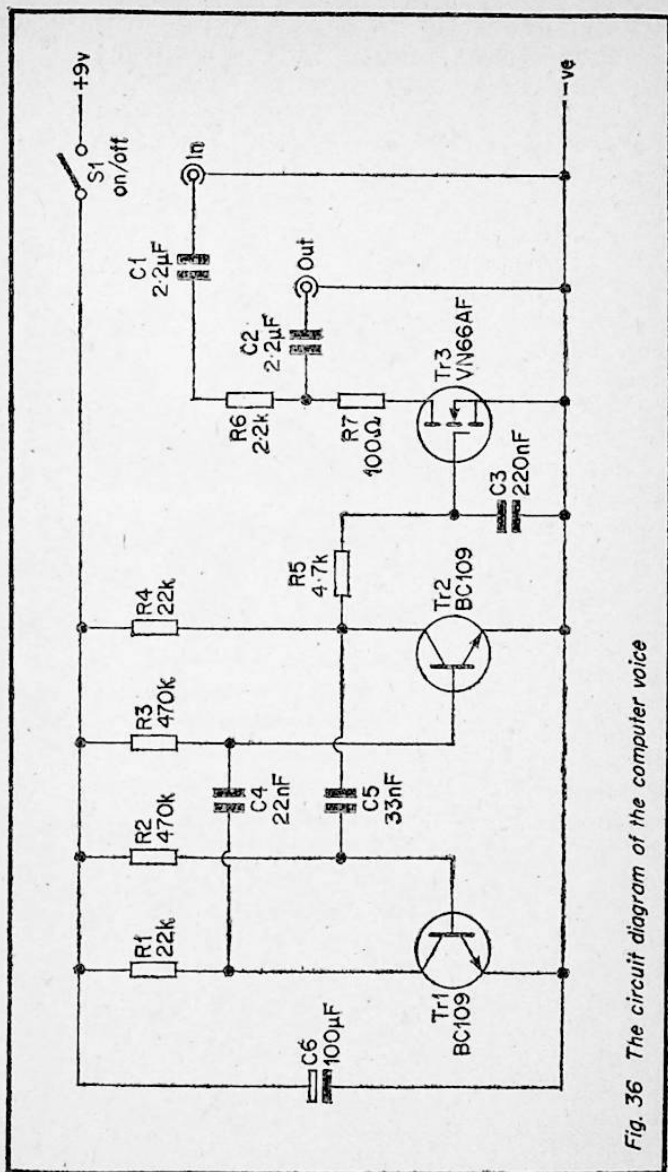


Fig. 36 The circuit diagram of the computer voice

Components: Computer Voice (Figure 36)

Resistors (all 1/3 watt 5%)

R1	22k	R5	4.7k
R2	470k	R6	2.2k
R3	470k	R7	100 ohms
R4	22k		

Capacitors

C1	2.2 μ F plastic foil	C4	22nF plastic foil
C2	2.2 μ F plastic foil	C5	33nF plastic foil
C3	220nF plastic foil	C6	100 μ F 10V

Semiconductors

Tr1	BC109	Tr3	VN66AF
Tr2	BC109		

Switch

S1 SPST toggle

Miscellaneous

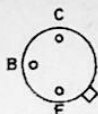
Case, component panel, 9 volt battery and connector to suit, sockets, wire, solder, etc.

CONCLUSIONS

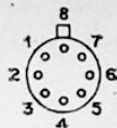
It has only been possible to include here some of the areas of application where VMOS and power MOSFET devices can be employed, and some of the more specialised areas such as receiving and transmitting RF amplifiers have had to be omitted. However, this should give the reader an insight into the versatility of these interesting and useful devices, and the ways in which they can be used.



VN10KM



BC107
BC109
BC109C
BC177
BC178
BC179



CA3130T
CA3140T



BD512
BD522

Heat-tab
connects
to drain



VN46AF
VN66AF
VN67AF
VN88AF

Case is
source

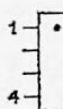


2SJ48
2SJ49
2SJ50
2SK133
2SK134
2SK135

Case is
drain



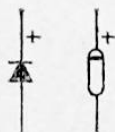
VN64GA



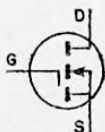
LF351
CA3130E
CA3140E



4001
4011



Diodes and
rectifiers



Symbol
n-channel device



Symbol
p-channel device

Fig. 37 Transistor base views, IC top views, and circuit symbols

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ISBN: 0 85934 027 9

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ISBN: 0 900162 70 8

256 Pages

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