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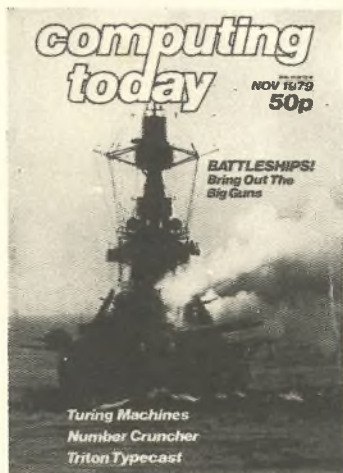
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Electronic Circuit Design

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INTERNATIONAL

No. 3

SPRING 1980

Introduction

This volume is the third in the present series of circuit design reprints from Electronics Today International. It will probably be a year or so before we are ready to publish Electronic Circuit Design Number Four.

The first volumes of this series have proved very popular: in conjunction with our "Top Projects" and "Circuits" reprints they provide the designer with a

wealth of information on today's techniques.

Publishing Electronic Circuit Design is an attempt to bridge the time gap between the information published in books and the up-to-the-minute editorial we publish in our monthly magazine, ETI.

Steve Braidwood,
Assistant Publisher — ETI

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DESIGNING AND USING ACTIVE FILTERS

A short series by Tim Orr which will enable the home constructor to utilise circuits of high complexity as easily as plugging in a resistor.

There is no doubt that active filters are very useful devices. Also, there is no shortage of literature on the subject. This would seem to suggest that designing active filters is a fairly straightforward business. Well, it is and it isn't. It is if you read *this* article. It isn't if you read the aforementioned literature. Most of the books on this subject have filled our heads with terms such as poles and zeros, Laplace transforms, transfer functions, etc, which haven't actually helped us to design anything!

Some basic theory

It is advisable to quickly run through some basic terms and expressions. Firstly, consider a simple low pass filter, Fig. 1a. The frequency response (Fig. 1b) is

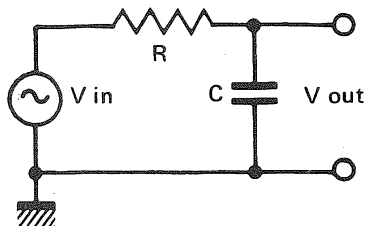


Fig. 1a. Simple low pass filter.

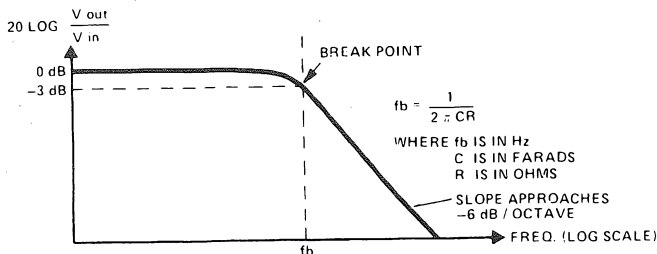


Fig. 1b. Frequency response of above.

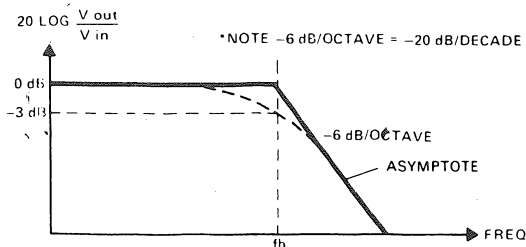


Fig. 1c. Approximation to response.

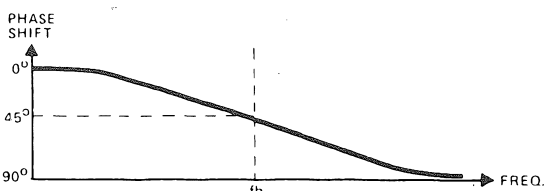


Fig. 1d. Phase shift v Frequency plot.

nearly flat until the break point, denoted by f_b . After this point the response rolls off at 6dB/octave, that is signals above this frequency are increasingly attenuated. The break point is defined as being the frequency where the resistance equals the capacitive reactance. When this occurs, the output is attenuated to 0.707 (-3dB) of the input. Although the resistance equals the capacitive reactance, the output is *not* half of the input. (This is because it is the vector sum of the two and hence equals 0.707 of the input!)

As the frequency response is a rather complex curve it is very useful to use a straight line approximation to it. These lines are called asymptotes (Fig. 1c). Note that the frequency response graph uses the convention of logarithmic scales, octaves or decades along the frequency axis, and dBs along the vertical axis representing output voltage divided by input voltage.

Phase shift with respect to frequency is also often plotted as in Fig. 1d. These two (the phase and frequency response plots) are known as Bode diagrams and are generally considered the most useful way of representing a filter's performance.

You will note that for the lowpass filter of Fig. 1a, the phase shift starts at 0°, is 45° at f_b and then approaches 90° as the frequency approaches infinity. This is not an *active* filter, it is composed entirely of *passive* components which means that its output cannot be effectively loaded without changing its performance.

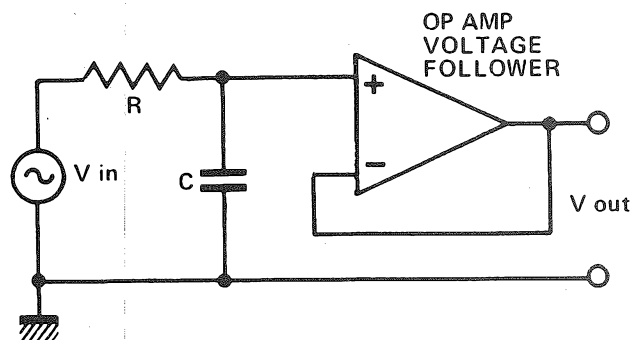


Fig. 1e. Active filter to perform the same task as the passive circuit of Fig. 1a.

Fig. 1e shows the same filter but in its active form, the op amp being used as a voltage follower serving only to isolate the filter's output. This type of filter is known as a First Order filter — a measure of the roll off slope.

When a more rapid slope is required, a higher order filter structure (one with more reactive elements) must be used. This is dealt with later.

Filter type	Low pass
Filter order	First order
Roll off slope	-6dB/octave or -20dB/decade (the same)
Breakpoint fb	$f_b = 1 / 2\pi CR$ Hz
Phase shift at fb	45°

TABLE 1. Summary of low pass filter of Fig. 1.

Passing highs

Next, let us consider the simple high pass filter of Fig. 2a. It is the complement of the low pass filter, the elements having been interchanged. Therefore it is not

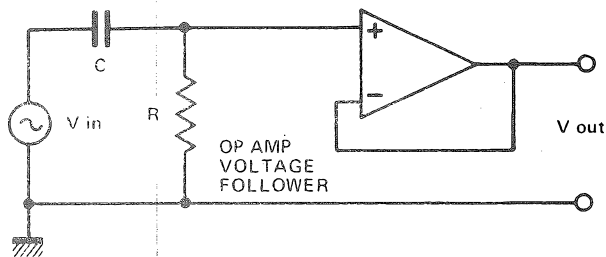


Fig. 2a. Simple high-pass active filter.

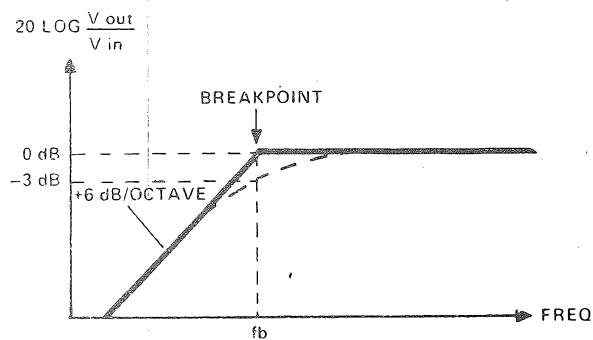


Fig. 2b. Frequency response of the high-pass filter.

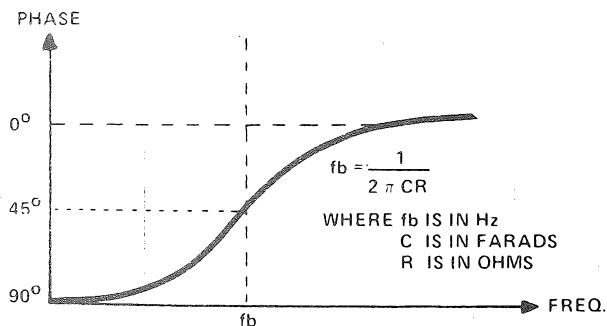


Fig. 2c. Phase v frequency plot of the same filter.

difficult to accept the complementary phase and frequency response curves of Fig. 2b. Note that the break point is the same and so is the roll off slope.

Filter type	High pass
Filter order	First order
Roll off slope	+6dB/octave or +20dB/decade
Break point fb	$f_b = 1 / 2\pi CR$ Hz
Phase shift at fb	45°

TABLE 2. Summary of the high pass filter Fig. 2.

Passing bands

The next type to be considered is a simple band pass filter shown in Fig. 3a. Although it uses an inductor it is only to illustrate the bandpass theory. Later on in this series, inductors will be replaced by their active equivalents.

The frequency response (Fig. 3b) shows that this circuit is symmetrical, having roll off slopes of 6dB/octave on either side of its RESONANT peak. This

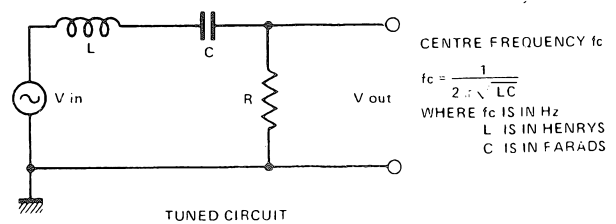


Fig. 3a. Simple band-pass filter.

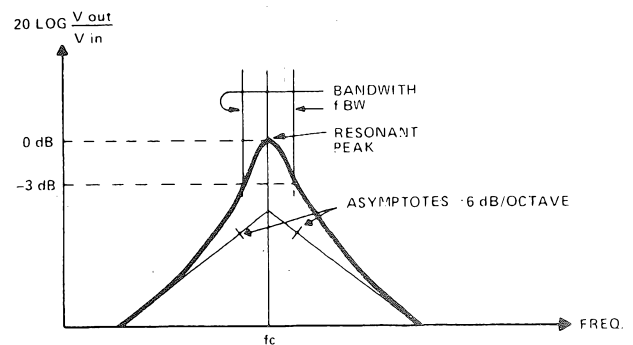


Fig. 3b. Band-pass frequency response.

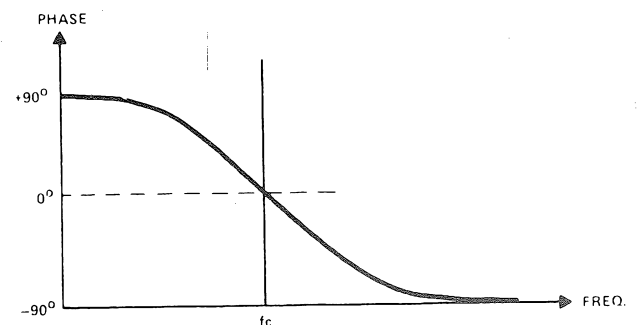


Fig. 3c. Band-pass phase response.

filter is known as a second order filter, because it has two reactive sections, the L and the C. The C is responsible for the +6dB/octave portion of the slope, the L for the -6dB/octave portion. But where these two slopes should meet, the response of the filter peaks and the slopes become much larger (Reson-

ance). The sharpness of this peak is described as the Quality of the filter, the Q factor. Resonance occurs at a frequency known as the Centre frequency denoted by f_c .

The bandpass filter is so called because it only passes signals within a certain bandwidth, which is defined as being the frequency range contained between the two points that are 3dB below the resonant peak. There is a fixed relationship between Centre frequency (f_c), bandwidth (fbw) and Q factor, given by $Q = f_c / \text{fbw}$.

The centre frequency is given by $f_c \approx 1 / 2\pi \sqrt{LC}$ Hz. This is only approximate, as it assumes that the value of R is relatively low. As R decreases, the Q factor increases. Thus R has the effect of damping the resonances, and so as it approaches zero ohms, Q approaches infinity.

The phase shift is shown in Fig. 3c. As this filter is a second order structure, then the total phase movement will be twice that for a first order structure, i.e. 180°. Fig. 3d shows the phase and frequency responses for different values of Q. Note that a high Q has a very rapid rate of change of phase, a low Q has only a slow rate of change.

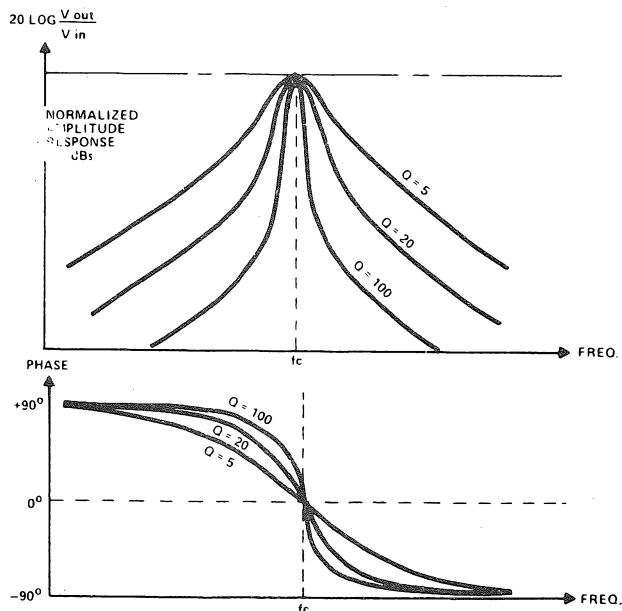


Fig. 3d. Effect of varying Q on the frequency and phase characteristics.

Time please

Bandpass filters also have a time response, as opposed to their frequency response. When an impulse is applied to a bandpass filter it rings (Fig. 3e). The filter oscillates at the centre frequency, f_c , the amplitude of the oscillations decaying exponentially in time. The ringing time, T_r , is the time taken for the oscillations to decay to 37% of their initial value. Ringing time is related to the Q and f_c by the following equation:

$$T_r = Q / 2\pi f_c$$

When a high Q filter has been constructed, it may prove difficult to measure its Q factor accurately due to the narrowness of its bandwidth. However, if the filter is made to ring, a reasonably accurate measurement of the Q can be obtained by measuring T_r and f_c .

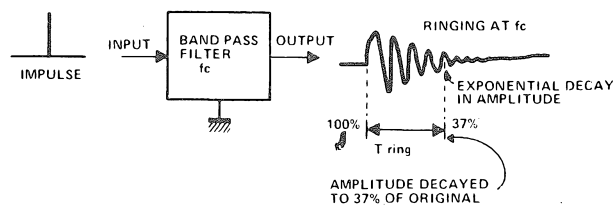


Fig. 3e. Ringing in a band-pass filter.

Filter type	Band pass
Filter order	Second order
Roll off slopes	+ and -6dB/octave greater near to resonance
Centre frequency f_c	$f_c \sim 1 / 2\pi \sqrt{LC}$
Phase shift at f_c	0
Q factor	f_c / fbw where fbw is the 3dB bandwidth
3dB bandwidth fbw	f_c / Q
Ringing time, T_r	$Q / 2\pi f_c$

TABLE 3. Summary of band-pass filter.

Failed band

Another common filter structure is the band reject or notch filter. There are many ways of realising this filter, one of which is shown in Fig. 4. The input signal is subtracted from the bandpass output. By adjusting R_a with respect to R, complete cancellation can be obtained at f_c .

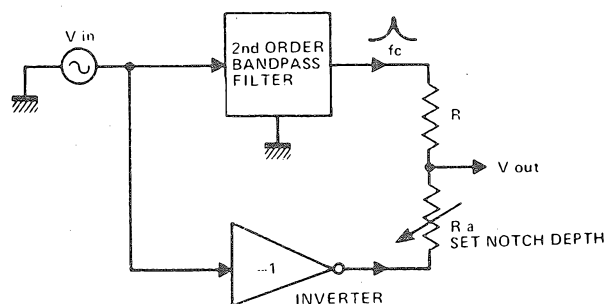


Fig. 4. Notch filter using Op-Amps.

Thus the centre frequency of the bandpass filter is the centre frequency of the notch, whose depth can be varied by altering R_a .

Very deep notches are possible, 50dB is easily obtained. As the Q of the bandpass filter is increased, so is the Q of the notch filter. However, R_a will have to be reset for each value of Q.

Filter Order

Consider the ideal low pass filter shown in Fig. 5a. Its response is flat right up until the break frequency f_b . Frequencies above f_b are attenuated to nothing! You won't be surprised to learn that filters like this don't exist. However, it is a common requirement to produce filters with very steep roll off slopes and this is achieved by designing filters with lots of sections. to increase the

filter order. Each reactive element in the filter increases the filter order by one, therefore a low pass active filter with three capacitors is known as a third order filter and will have an ultimate roll off of three times 6dB/octave, which is 18dB/octave.

However, designing a third order lowpass filter is not just a simple case of sticking three first order RC circuits in a line. What you get when you do this is a very soggy curve indeed! The filter should be flat in the pass band, then it should turn over and rapidly assume its ultimate roll off slope. Examples of this type of Maximally flat filter are shown in Figs 5b and c. The effect of order number upon a bandpass filter is shown in Fig. 5d.

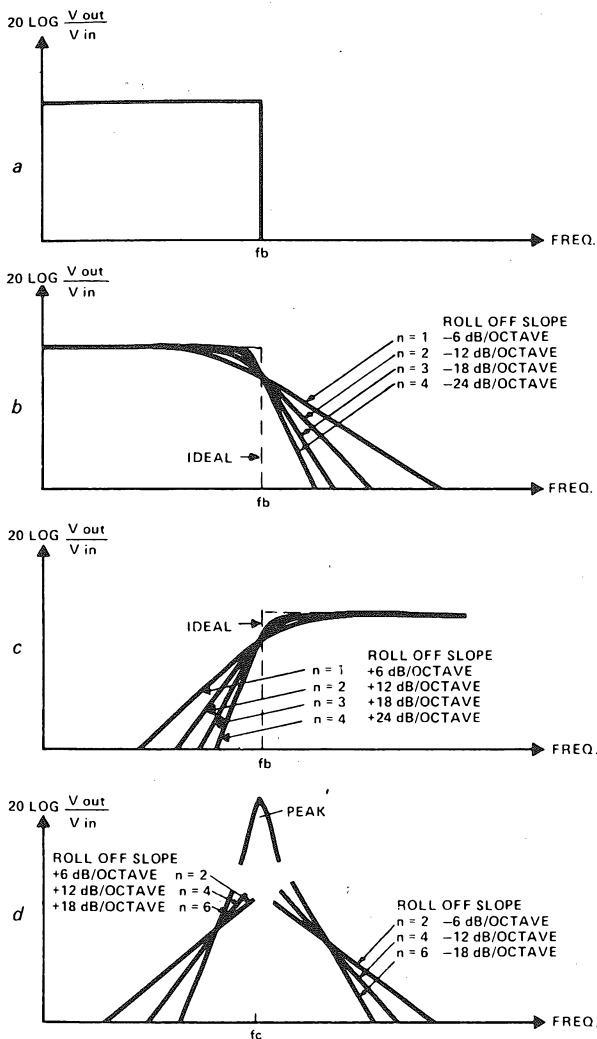


Fig. 5a. Ideal low-pass response:
b, c. Examples of maximally flat filter responses.
d. Effect upon band-pass filter of increasing order numbers.

Later on in this series the circuit diagrams and design charts are given for various filter types and order numbers. It would seem that to get a filter to approach its ideal response, all that is needed is to increase the order number. This is in fact true, but there are certain tolerance problems. (When 8th order filters are designed, component tolerances of about 1% are required!)

Which filter shape?

The type of filter that is chosen to do a particular job will depend on what parameters are thought to be important. There are three basic characteristics to be considered (lowpass and highpass filters only).

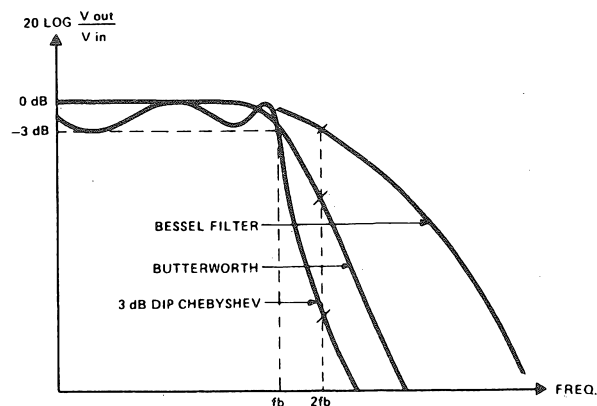
1. Good transient response.
2. Maximum flatness of the filter within its passband.
3. Maximum rolloff slope outside the passband.

The type of filter used should be chosen to fit the job that they are being designed for. The filters have been categorised into three basic types for the purpose of simplicity.

Filter number 1 is known as a Bessel filter. Its phase changes almost linearly with frequency. It is useful for systems where a good transient response is required, such as joining the dots up on the output of a digital to analogue converter. It has a very poor initial roll off slope.

Filter number 2 is known as a Butterworth filter. It has the flattest pass band possible. Its other two parameters are a compromise. That is it has a reasonable overshoot and a fairly fast initial roll off.

Filter number 3 is known as a Chebyshev filter. It has some ripple in its pass band, although this is small, and a very fast initial roll off, and a poor transient response.



ATTENUATION AT FIRST OCTAVE (2 fb)

*3 dB CHEBYSHEV	17	28	39	51	62	75
BUTTERWORTH	12	18	24	30	36	42
η /FILTER ORDER	2	3	4	5	6	7

*NOTE THE IMPROVED ATTENUATION

Fig. 6. Response of all three types of filter discussed, with table showing variation of attenuation between them.

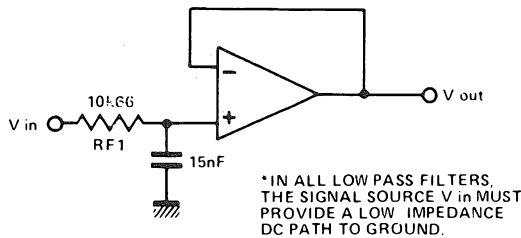
DESIGNING AND USING ACTIVE FILTERS PART 2

Continuing Tim Orr's instructive series designed to help the home constructor employ one of the most useful circuit blocks available

The following section contains all the information needed to be able to build low and high pass filters, of first, second, third and fourth order to Bessel, Butterworth and Chebyshev characteristics.

Low pass

Figure 1 shows a first order low pass filter. In all the examples to follow the filters have been designed for 1kHz operation. Equal component value Sallen and



*IN ALL LOW PASS FILTERS, THE SIGNAL SOURCE V_{in} MUST PROVIDE A LOW IMPEDANCE DC PATH TO GROUND.

	RF1	GAIN IN dB	COMPONENT TOLERANCES
BESSEL	10k66	0	10%
BUTTERWORTH	10k66	0	10%
CHEBYSHEV	10k66	0	10%

ALL THE RESPONSES ARE THE SAME FOR A FIRST ORDER FILTER

Fig. 1. A general circuit for a first order low pass filter.

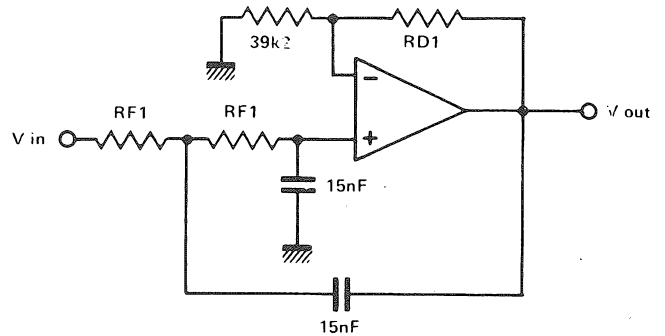
Key filters have been used as the basic building blocks. If operation at a frequency other than 1kHz is required, then the resistor/s R_f should be scaled accordingly, (the R_d resistors are not altered). For example, if operation is required at 250Hz, then the R_f in the chart must be multiplied by

$$\frac{1000}{250}$$

which is $\frac{\text{(Normalised 1kHz)}}{\text{(Required frequency of operation)}} = 4$.

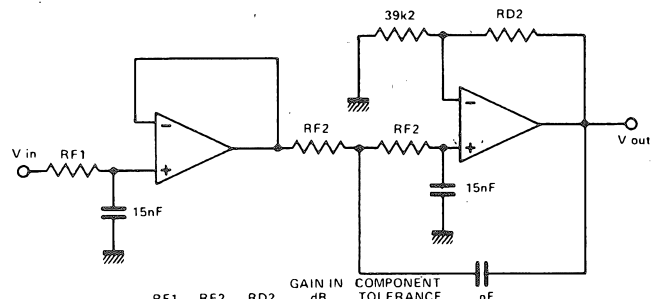
Figure 2 shows second, third and fourth order filters. The design procedure is as follows:--

1. Decide which type of filter is required, high, low, bandpass or notch.
2. In the case of high or low pass, decide which type of response is required, Bessel, Butterworth or Chebyshev.
3. Next, what filter order is needed. This will have led you to a particular order filter with components designed for 1kHz operation.
4. Scale the R_f components so that the filter will operate at the required frequency.
5. Build and test the filter.



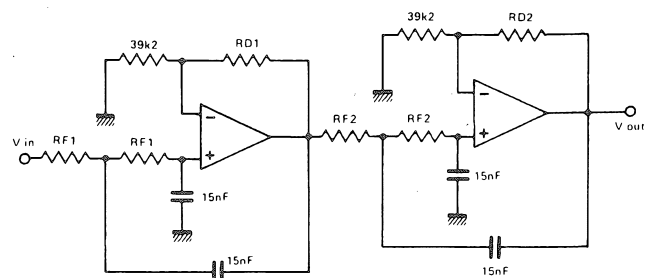
	RF1	RD1'	GAIN IN dB	COMPONENT TOLERANCE
BESSEL	8k39	10k5	2.3	10%
BUTTERWORTH	10k66	22k6	4.1	10%
CHEBYSHEV	12k6	48k7	6.8	5%

Fig. 2a Second Order low pass filter design, break frequency = 1kHz.



	RF1	RF2	RD2	GAIN IN dB	COMPONENT TOLERANCE
BESSEL	8k	7k26	21k5	4.1	10%
BUTTERWORTH	10k66	10k66	39k2	6.0	10%
CHEBYSHEV	35k41	11k73	66k5	8.6	2%

Fig. 2b. Third Order low-pass filter. To alter break frequency (here 1kHz) scale resistors accordingly.



	RF1	RD1	RF2	RD2	GAIN IN dB	COMPONENT TOLERANCE
BESSEL	7k45	3k24	6k60	29k4	5.6	10%
BUTTERWORTH	10k66	5k9	10k66	48k7	8.3	5%
CHEBYSHEV	24k11	42k2	11k20	71k5	15.3	1%

Fig. 2c Fourth Order Low Pass Filter.

There are of course some problems which may occur. One is that these filters have a voltage gain in their passband. So you might find that although you have got the required frequency response there is an unexpected signal gain.

This may cause some problems with op-amp bandwidth. As a rule of thumb, the op amps should have 10 to 100 times more bandwidth than the product of the filter's maximum operating frequency times the individual stage gain of each section. If the op amp runs out of bandwidth or introduces a phase shift then the filter is not going to work properly. For the examples given, if you use a 741 as the op amp then a frequency limit of approximately 10kHz should be imposed. (If an LM318 is used then the limit can go to 200kHz). Another problem is one of range of values of R_f . If R_f is made too small then large currents have to flow from the Op amp and this may effect the performance of the filter. If R_f is too large there may be hum pick-up problems and DC offset voltage problems due to bias currents. Therefore, keep R_f between 1k and 100k. If R_f needs to exceed this range, scale the capacitor as well.

Charting examples

As an example of using the design tables, let us solve the following problem. Design an audio 'scratch' filter, having a break frequency of 7.5kHz and an attenuation at 15kHz of more than 20dB. The first decision to be made is what type of response do we want? A roll off of more than 20dB/octave is quite steep and so the Bessel filter is ruled out. The Chebyshev filter has a poor transient response and at 7.5kHz we would hear it ringing. Therefore a Butterworth response should be used. Next, the filter order. Third order gives us -18dB/octave which is not sufficient, fourth order gives -24dB/octave . Hence what is needed is a fourth order Butterworth design (fig. 2c).

The break frequency is 7.5kHz and so the resistors R_{f1} and R_{f2} have to be *divided* by 7.5. This gives $R_{f1} = 1\text{k}42$, $R_{f2} = 1\text{k}42$, $R_{d1} = 5\text{k}9$, $R_{d2} = 48\text{k}7$, $C = 15\text{nF}$, and the component tolerance is 5%. Now we must fit preferred values to the resistors.

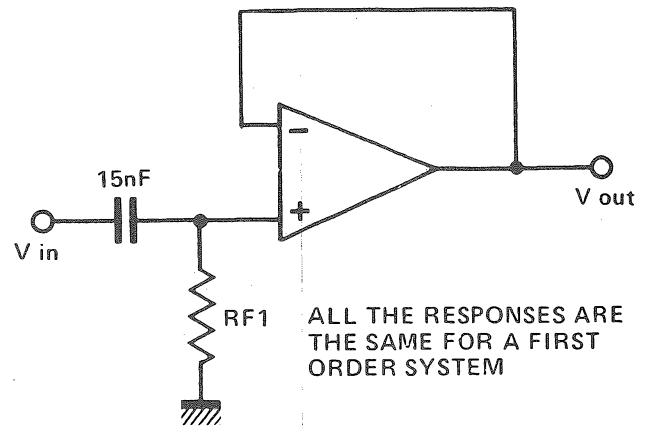
R_{d2} becomes 47k, R_{d1} becomes 6k2 (this is just over the limit of tolerance) R_{f1} and R_{f2} are a problem. Even when taken to the nearest E24 value they are outside the component tolerance allowed. There are two solutions: use the nearest E96 1% resistor or use 1k5. This will lower the break frequency by about 6%, but as this is only an audio filter no one will probably be any the wiser!

High Pass

Figure 3 gives the design tables for high pass filters. The design procedure is exactly the same as that for low pass filters.

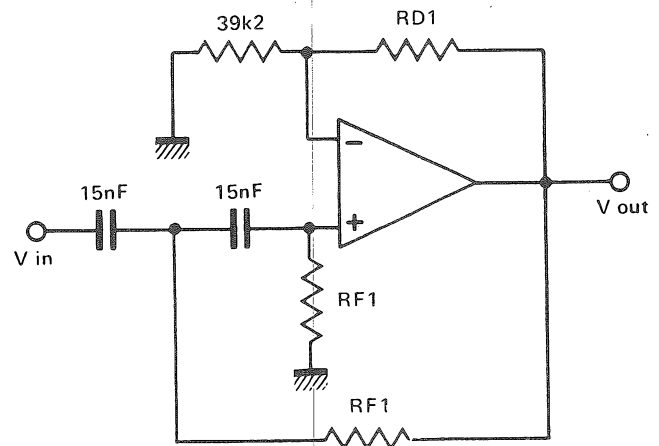
Band Pass

Several second order band pass filters can be cascaded to produce a different response shape which, like those discussed earlier for the low and high pass filters, can be optimised to give maximum roll off, or maximum pass band 'flatness'. However, these tend to get rather difficult to design and so only second order filters will be discussed.

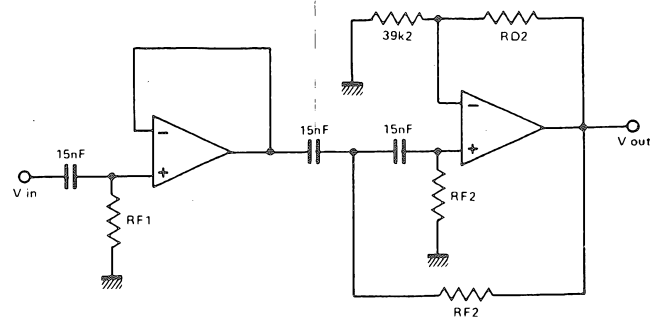


ALL THE RESPONSES ARE THE SAME FOR A FIRST ORDER SYSTEM

	R_{f1}	GAIN IN dB	COMPONENT TOLERANCE
BESSEL	10k66	0	10%
BUTTERWORTH	10k66	0	10%
CHEBYSHEV	10k66	0	10%



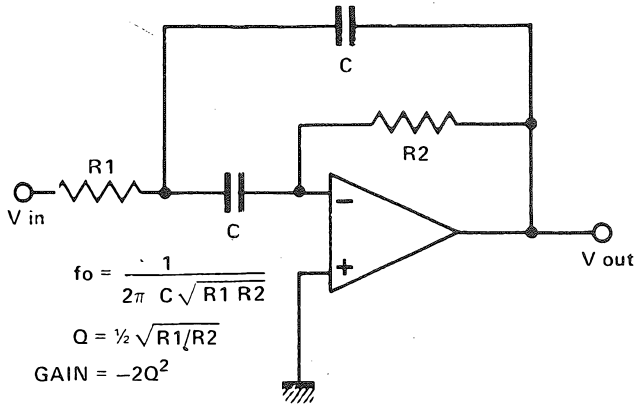
	R_{f1}	R_{D1}	GAIN IN dB	COMPONENT TOLERANCE
BESSEL	13k55	10k5	1.3	10%
BUTTERWORTH	10k66	22k6	1.6	10%
CHEBYSHEV	9k01	48k7	2.2	5%



	R_{f1}	R_{f2}	R_{D2}	GAIN IN dB	COMPONENT TOLERANCE
BESSEL	14k19	15k68	21k5	4.1	10%
BUTTERWORTH	10k66	10k66	39k2	6.0	10%
CHEBYSHEV	3k21	9k70	66k5	8.6	2%

Fig. 3. From the top! First, second and third order high pass filters, break point 1kHz. Final roll off is 6, 12 and 18 dB/octave respectively.

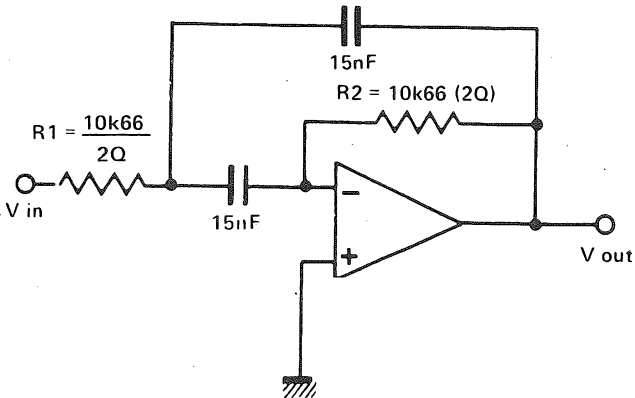
Figure 4 shows a simple bandpass filter known as a multiple feedback circuit. This circuit can only provide low values of Q up to about 5. It will probably oscillate if it is designed to give a higher Q. Note that a high Q implies a large gain at the centre frequency. Therefore care must be taken to ensure the op amp has enough bandwidth to cope with the situation. Fig. 4



$$f_0 = \frac{1}{2\pi C \sqrt{R_1 R_2}}$$

$$Q = \frac{1}{2} \sqrt{R_1/R_2}$$

$$\text{GAIN} = -2Q^2$$



Q	R1	R2	GAIN IN dB
1	5k33	21k32	6 dB
2	2k66	42k66	18.1 dB
3	1k77	60k40	25.1 dB
4	1k33	85k33	30.1 dB
5	1k06	106k66	34.0 dB

Fig. 4. A multiple feedback bandpass filter. The centre circuit is normalised for 1kHz. The table is the design table for this circuit. To change the centre frequency change R₁ and R₂ by an equal factor.

gives a design chart, normalised for 1kHz operation. First, choose a Q factor and then perform the frequency scaling. For instance, if the centre is 250Hz, then multiply both R₁ and R₂ by a factor of 4. If a high Q is required, then a multiple op amp circuit must be used. The 'state variable' and the 'Bi-Quad' are two such circuits and Q's as high as 500 may be obtained with them.

Figure 5 shows a state variable filter. It has three major features which are

1. It can provide a stable high Q performance.
2. It is easily tuned.
3. It is versatile, providing bandpass, lowpass and highpass outputs simultaneously.

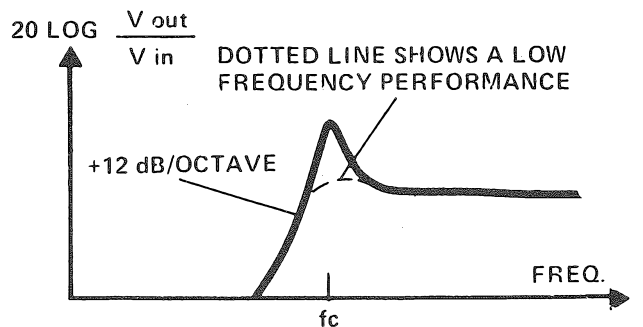
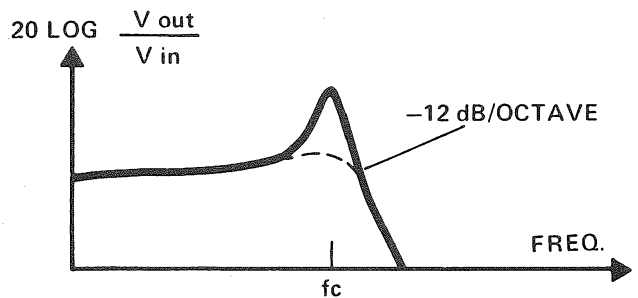
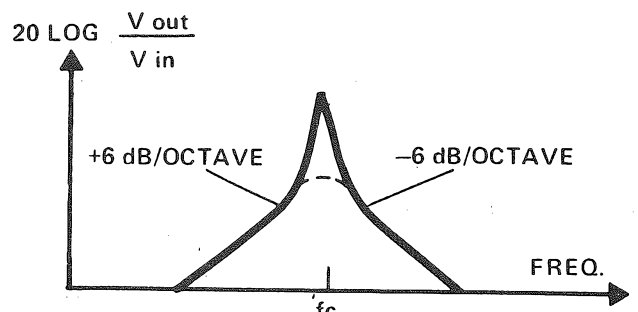
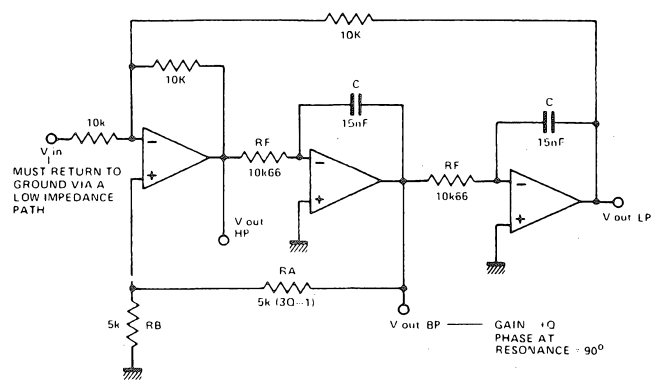


Fig. 5. The state variable filter is called a universal filter because it can give bandpass, low and high pass outputs — as shown above. Note that all these responses are second order in nature.

The Q is determined by the ratio of two resistors, R_A and R_B where R_A/R_B = 3Q - 1. The resonant frequency f_c =

$$\frac{1}{2\pi RC}$$

Note that there are two C's and two Rf's in the circuit, and so if the filter is to be tuneable, then both Rf's should change by an equal amount (the Rf's can be a stereo pot).

You will note that Q and fc are independent of each other, and so as the resonant frequency is changed, Q remains constant, and vice versa.

Op amps

The requirements placed upon the op amps in the filter, Fig. 5, are less than that for the multiple feedback circuit. The op amps need only have an open loop gain of 3Q at the resonant frequency. Say we have a Q of 100 and an fc of 10kHz. Therefore the open loop gain is 300, the frequency is 10kHz and so the gain bandwidth product needed is 3MHz. When using a high Q, care must be taken with signal levels. The gain of the filter is +Q at resonance, and so if you are filtering a 1V signal with a Q of 100 then you could expect to get a 100V output signal!

National Semiconductors manufacture an active filter integrated circuit, which is a four amp network that can be used to realise state variable filters with Q's up to 500, and frequencies up to 10kHz. The device is called AF100.

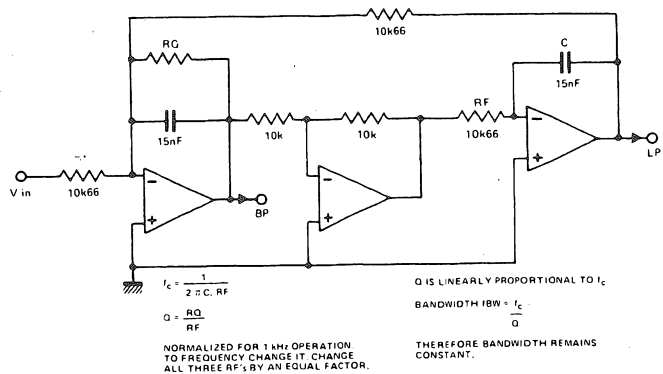


Fig. 6. A Bi-Quad active filter design.

Figure 6 shows a Bi-Quad active filter. It looks very similar to the state variable filter, but the small changes make it behave quite differently. It only has a bandpass and a low pass output. The resonant frequency is given by

$$f_c = \frac{1}{2\pi CR_1}$$

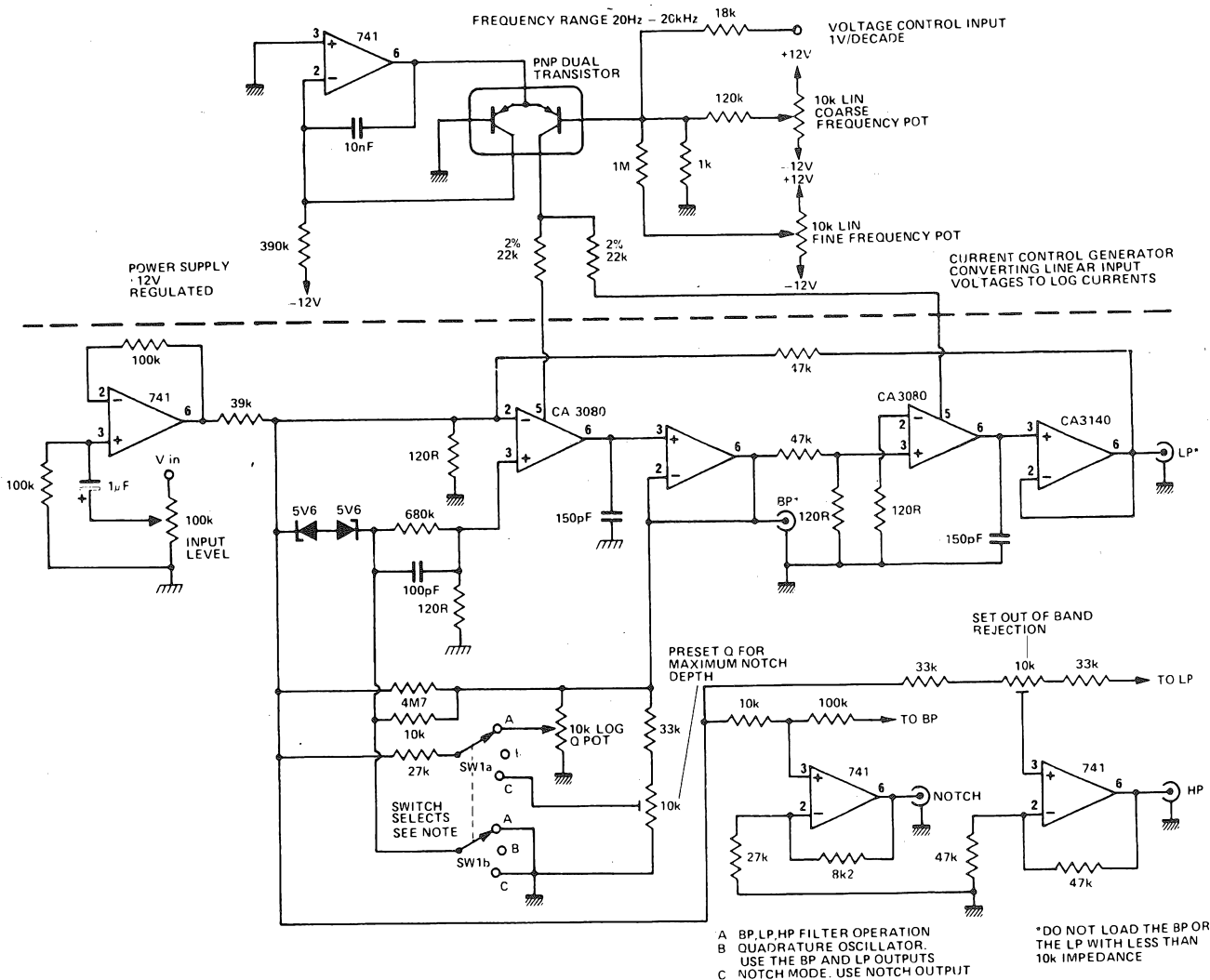


Fig. 7. The state variable filter can also be made to oscillate (as above). It has a variable resonant frequency, it becomes a variable frequency oscillator. This circuit produces two low distortion sinusoids in phase quadrature; ie, sine and cosine waveforms at low distortion.

DESIGNING AND USING ACTIVE FILTERS PART 3

Concluding our detailed examination of this particular building block Tim Orr takes a good look at band pass and band reject circuits.

Band reject (notch) filters

So far notch filters have been realised in this article by two methods; by mixing a bandpass signal with the original or by mixing the low and high pass outputs. There are of course, many other methods of obtaining a notch.

Firstly, the Twin T circuit, Fig. 1.

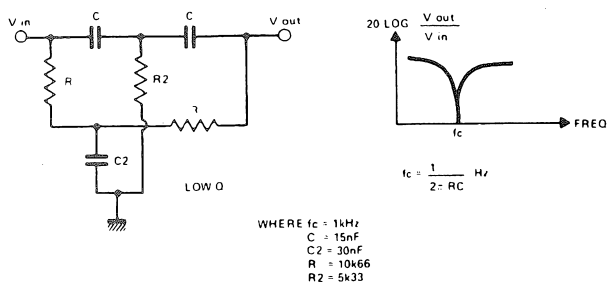


Fig. 1. Twin-T band notch filter configuration.

This is interesting, in as much as by using only resistors and capacitors, a notch response can be obtained! However, as this filter is passive, only a low Q is possible. This circuit is not used very much, because it has six components that determine its notch frequency. However, it is of interest to note that, when the Twin T is placed in the feedback loop of a high gain inverting amplifier, a bandpass response is obtained. Also if R is made variable it is possible to move the centre frequency, although in doing so, the Q varies. This has been the basis of many Wah Wah effects pedals, Fig. 2.

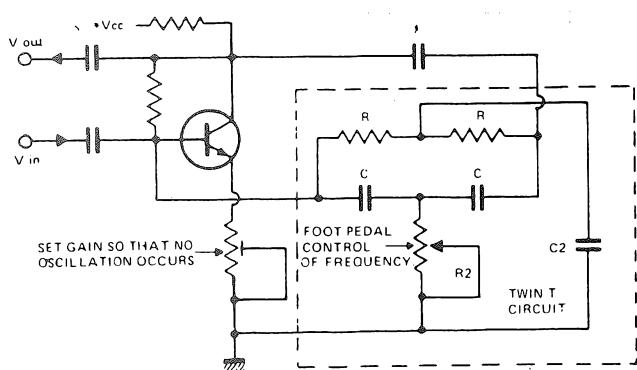


Fig. 2. Block Diagram of a typical Waa-Waa Pedal.

Another method of obtaining a notch is to use the 'Allpass' filter, Fig. 3. The frequency response shows that its output is flat! Not much of a filter I hear you saying. However, it suffers a phase shift which goes from 180°, through 90° at f_c, to 0. By cascading two of these filters, the phase shift is doubled.

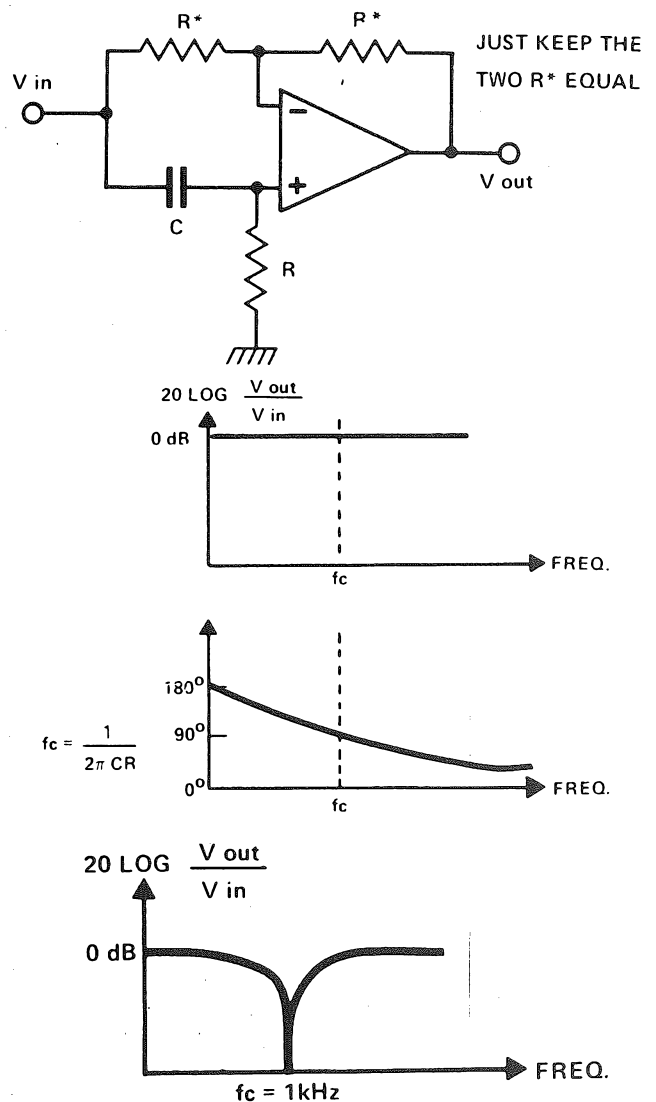


Fig. 3. All-Pass filter. At the top is the circuit for such a device. Its frequency and phase response are shown below it, with the obtainable notch at the bottom.

If we then mix the phase delayed signal with the original, a notch response is obtained. This is because at ω the two signals have the same magnitude, but the opposite phase and so they cancel each other out.

If the notch is to be made tuneable, then the RC time constants must be varied. For a small tuning range just one R can be varied, for a large tuning range then the R's must be realised with a 'stereo' pot.

All change

If lots of Allpass filters are cascaded then several notches can be produced. This type of filter is known as a comb filter. Note that it takes two Allpass filters to produce a usable 180° phase shift, and therefore every notch in the comb requires two Allpass sections. By making the R's variable then the notches can be made to move up and down in frequency. This filter forms the well known 'phasing' effect unit, widely used in the music industry to produce colouration!

Fig. 4 shows a small section of just such a unit. A CMOS chip is used to provide a MATCHED set of six MOS FETS. A common voltage is used to control the MOS FETS channel resistance. Thus as the control voltage varies then so do the six MOSFET resistors, and the three notches move in unison along the frequency axis.

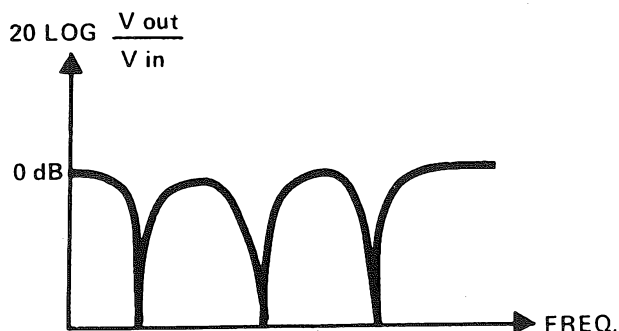
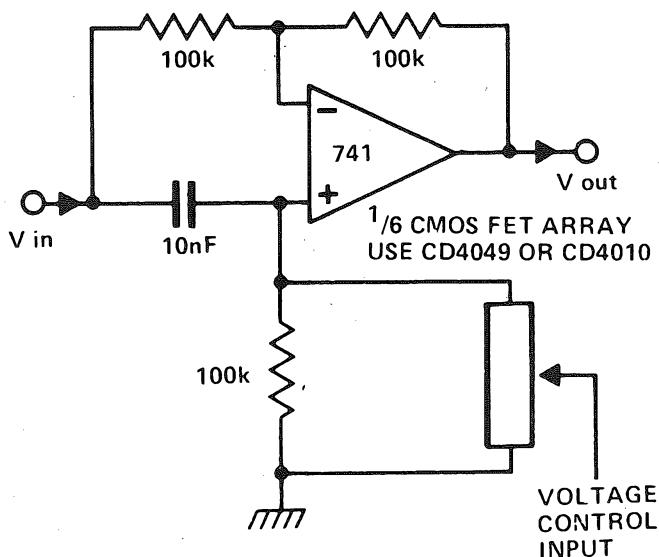


Fig. 4. One section of a comb filter. The response produced by the full (six times above) circuit is shown below the circuit.

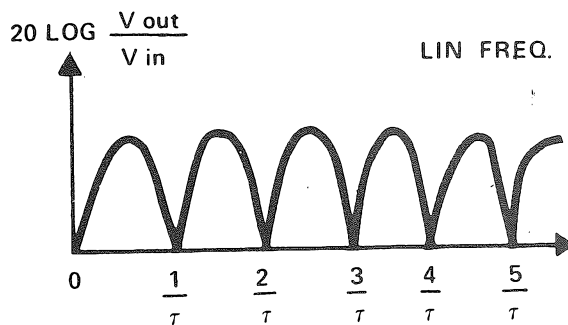
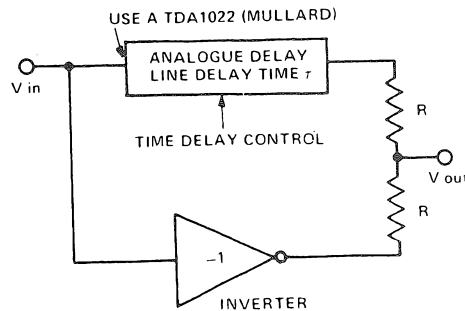


Fig. 5. Alternative method of producing a comb filter using a Mullard delay line.

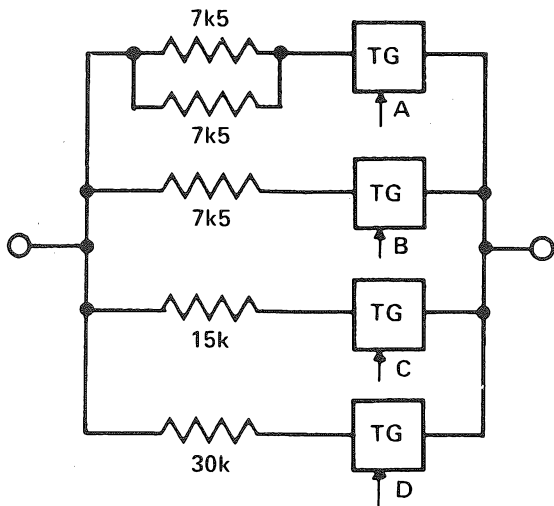
Another form of comb filter is shown in Fig. 5. Instead of a phase delay line, a time delay line is used. This produces a large number of notches which are linearly spread along the frequency axis, their spacing being determined by the delay time.

A bucket brigade delay line can be used to implement the time delay and this can be made variable. This type of filter is known as a Flanger, which is a superior type of phasing unit, and is used to generate high quality phasing effects. An even more impressive sound can be produced by adding some feedback around the delay line. A multi peak, high Q filter is formed which makes very interesting musical sounds when swept.

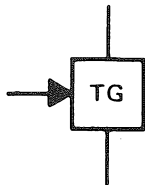
Variable Tuning

Very often a variable centre or cut off frequency is wanted. This causes problems in filters of order greater than two, simply because getting ganged potentiometers with more than two sections is difficult. One well known manufacturer uses four presets mounted on a common spindle to produce a fourth order Rumble and scratch filter. For manually controlled filters, the resistors are made variable by using ganged potentiometers or switched resistor networks.

The switches can be mechanically operated or electronically controlled, Fig. 6. An alternative method of switching control is to use mark/space modulation, Fig. 7. This has the advantage of being a continuously variable control with a useable sweep range of 100 to 1. Also, lots of sections can be used, and they will all track. Therefore, if two CD4016 packs were used (4 transmission gates per pack), then an eighth order variable frequency filter could be constructed. There are, of course, problems.



CONTROL VOLTAGES A,B,C,D PROVIDE A RANGE OF 15 VALUES OF RESISTANCE.



CMOS CD4016 or CD4066

Fig. 6: Varying the tuning of an active filter by use of CD4016 transmission gates (TG) to switch in different resistor values.

1. The switching waveform must be several times higher in frequency than the highest frequency to be filtered.
2. More circuitry, to generate the switching waveform is required.
3. Switching noise is generated.

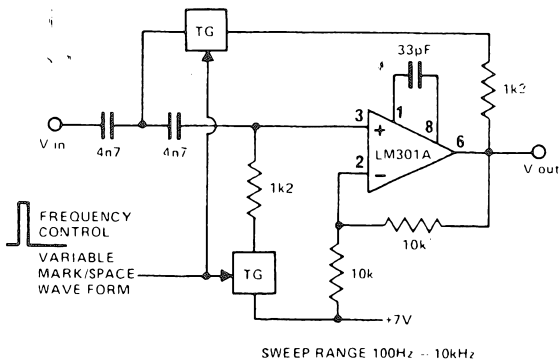


Fig. 7: Another method of varying the notch frequency, mark / space ratio modulation, has the advantage of possessing a wide range.

Multiplying FETS

Voltage controlled resistors can be used. These take the form of junction or MOS FETS, where the gate voltage controls the channel resistance, R_{ds} . The problems with this method are that the characteristics from FET to FET vary considerably and also the R_{ds} does not have a predictable relationship to the gate voltage. Also, to avoid distortion, low signal levels must be used. Nevertheless, FETS are used in many variable filters such as phasing units.

A set of six MOS FETS having matched characteristics can be obtained from a CD4049 or a CD4010 pack. Alternatively LED photo conductor arrays can be used. The LED produces light which controls the photo conductor's resistance, the two devices being housed in a lightproof box. Large signals can be handled with very low distortion and low noise generation.

Again there are drawbacks. The units are quite expensive, the relationship between LED current and photo conductor resistance is rather unpredictable and the photo conductor's characteristics drift. Another method of varying a filter frequency is to use electronic multipliers. A two quadrant multiplier function can be used to vary the gain of a stage and so produce frequency scaling.

Some Audio Circuits

Active filters have found great use in equalising audio signals, from tone controls on a domestic Hi-Fi to parametric equalisers in recording studios. Fig. 8

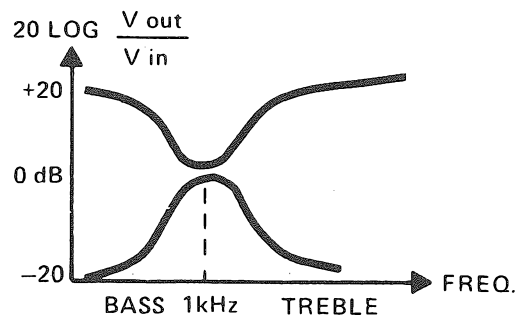
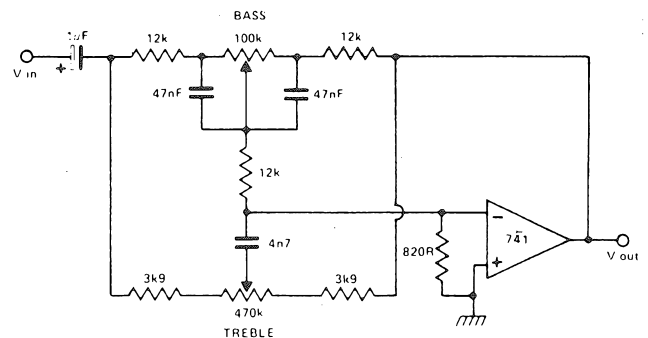
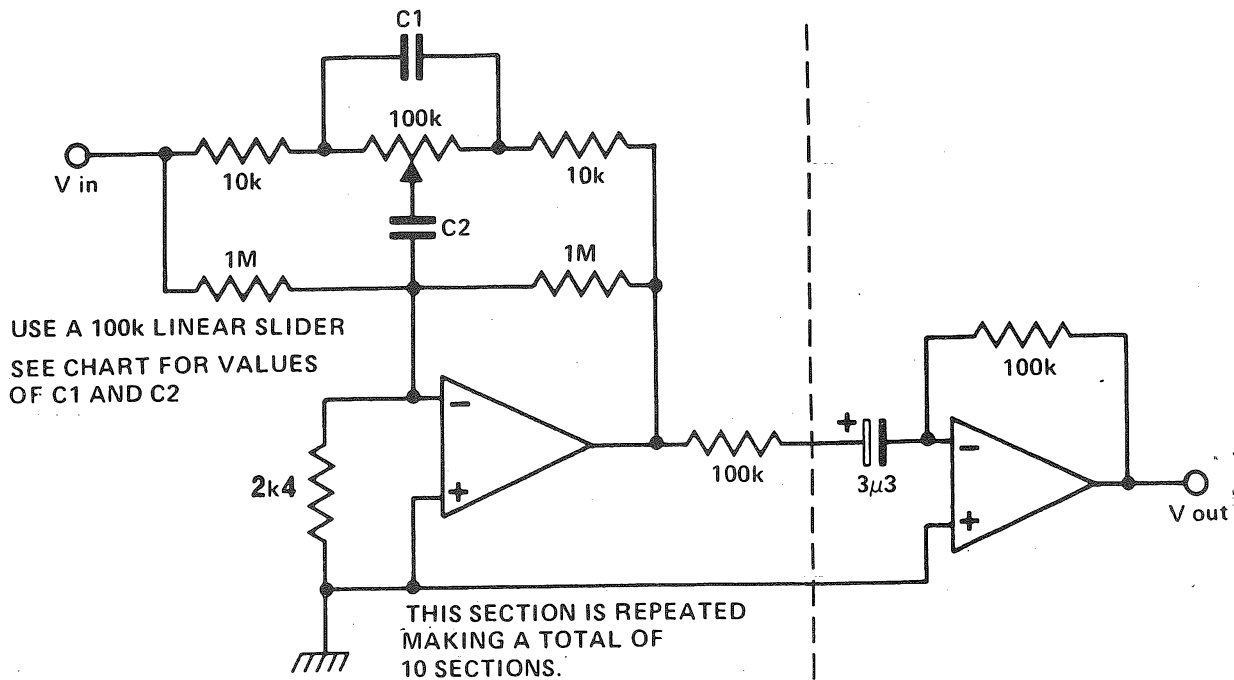
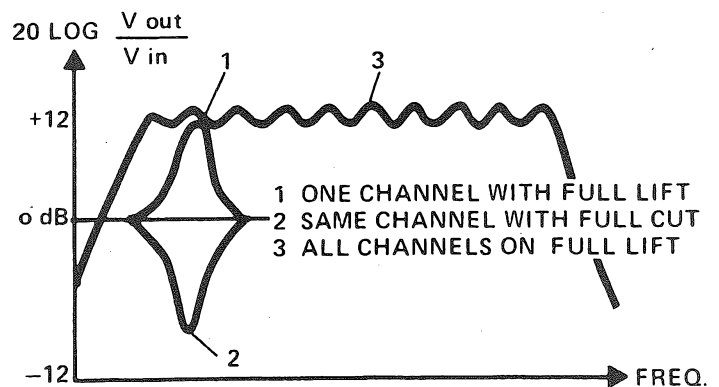


Fig. 8: Simple tone control circuit, with the lift and cut responses shown beneath it.



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

Fig. 9: A design for a graphic equaliser, using active filters. The above circuit is repeated however many times you wish. Use the table on the left to calculate values.



shows a common tone control with just bass and treble functions. Cut and lift ranges are 20dBs. If a more flexible control of the spectrum is needed then a ten band graphic equaliser (Fig. 9) could come in handy.

Testing Designs

Once the process of designing active filters has been reduced to a simple procedure, testing them should also be made as easy as possible. The most basic is to use a swept sine wave oscillator (Fig. 10).

An XY oscilloscope is used to display amplitude (linear) against frequency (log). The ideal display would be log amplitude, but this is not so easy to obtain. The beauty of this method of testing is that the display is real time and so any changes made to the filter, like varying one of the capacitors, appear instantly on the oscilloscope. If high Q's or rapid roll offs at low frequencies are involved then the sweep time will have to be reduced, otherwise the effects of Ringing, will 'Time smear' the display. The harmonic distortion of the sine wave can be quite large, 0.5 to 2.0% without causing too much of a display problem for most filter designs.

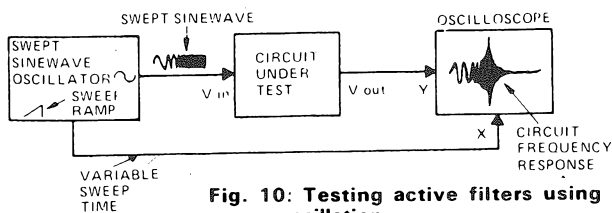


Fig. 10: Testing active filters using a sweep oscillation.

A FEW CHEAP TRICKS!

Want a stable low voltage? Want to fire a thyristor without using unijunctions, or even make a thyristor? Whatever your semiconductor problem there is probably a cheap way round it.

When you look over all the circuits that are published in the time of one month, you might imagine you'd need several rooms just to hold all the semiconductors that are needed. It's not really so and the cunning experimenter can use several dodges to get by with a very limited stock indeed. There are several project designers, for example, who manage to test out their ideas using no more than two transistor types, a 2N2219 and a 2N2905. These are silicon switching transistors which look exactly alike and differ only in polarity — the 2219 is NPN and the 2905 is PNP. How's it done? Read on.

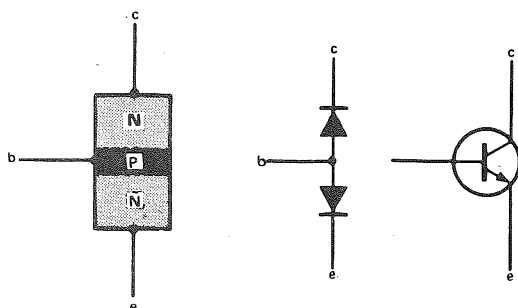


Fig. 1. Structure of a transistor. (a) The semiconductor sandwich, (b) connection of two diodes which gives the same readings when connected to resistance meters, (c) symbol (NPN illustrated).

Basically, a transistor is constructed like two back-to-back diodes (Fig. 1), the difference being that both diodes form part of one crystal. We can, therefore, use a transistor to substitute as a diode. Which bit do we use? The collector and base terminals form one diode, a high reverse voltage diode which will pass quite large currents. Transistors of the 2N2219 variety will dissipate 0.8W at the collector, so that their collector base diodes can be quite happily used in bridge rectifier circuits for up to 30 V supplies, keeping the emitter open circuit or shorted to the base.

A Bit Of Bias

The base-emitter diode, on the other hand, is much more of a small signal diode, more suited to low current, low

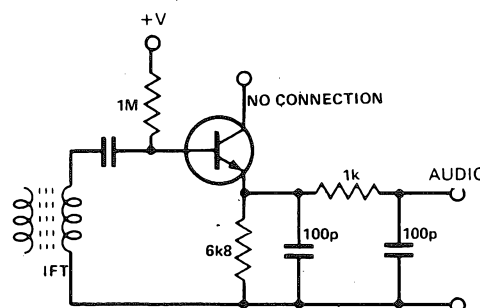


Fig. 2. Using the base/emitter junction of a transistor as a detector diode. The 1M resistor keeps the junction slightly conducting, so increasing the sensitivity.

voltage work. One minor drawback is that you can't approach the small forward voltage of a germanium diode, but there's no law to say you can't apply a bit of bias, as in Fig. 2. This makes the base emitter diode into a good, sensitive detector. While we're on the subject of detectors, why not be different and use an emitter follower detector, as in Fig. 3? It's a darn sight more linear than a straightforward diode, and has a low output impedance and high input impedance as well.

The circuit is a simple one. A capacitor is connected across the emitter resistor of an emitter follower. The size of the capacitor should be such that the time constant of emitter resistor x capacitor is small compared to the time of an audio wave but large compared to the time of the

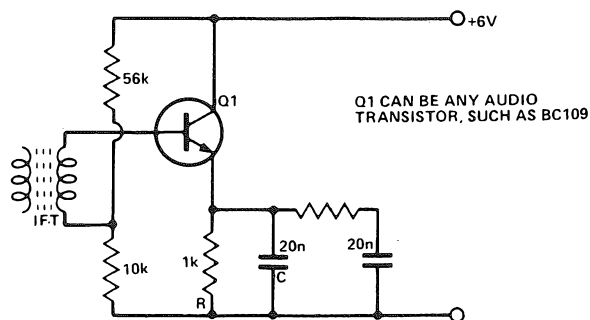


Fig. 3. The emitter-follower detector.

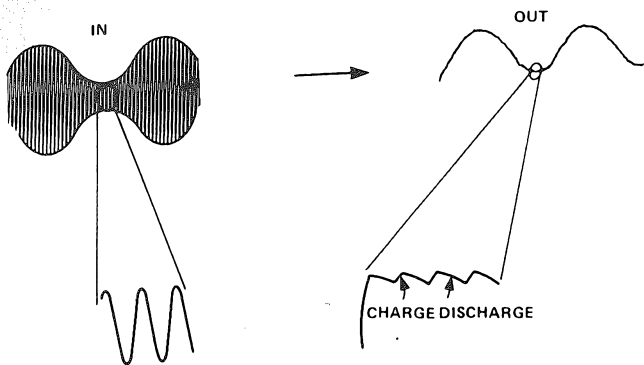


Fig. 4. Action of the emitter-follower detector. Capacitor C is charged by the current through Q1 during the positive part of a cycle, but can discharge only slowly through R. The voltage across C follows audio frequency changes, but not radio frequency changes.

RF wave. Time constants of 10 to 100 μ S are usually suitable for AM radio circuits, so that a typical circuit might use 1k emitter resistance and 20n (that's 0.02 μ) capacitance. The action is also straightforward (Fig. 4). The positive RF wave makes the transistor conduct, so that C1 charges up to the positive peak of the wave. Because the time constant is large compared to the time of one RF wave, though, the voltage at the emitter drops only slightly as the wave goes through the remainder of its cycle and the transistor cuts off until around the peak of the next RF wave. The AF modulation, however, makes the peaks of the RF signal occur at different voltages, tracing out the audio waveform, so that the audio signal appears at the emitter, with very little trace of RF so that nothing much in the way of filtering is needed. The emitter-follower detector also has lower distortion than the conventional diode detector.

Transistor Zener

We're not finished with diodes, though. The base-emitter diode of most planar silicon transistors (and that means most 'modern' silicon transistors manufactured in the last 15 years) will act as a zener diode. The circuit of Fig. 5 shows how this can be checked. The voltage across the base-emitter junction will stabilise at anything from 7 V to 18 V, depending on the construction of the transistor, when power is applied. You don't need to keep a drawer full of zener diodes, just make these 2N2219's work for their living.

This zener diode action, incidentally, can cause some odd effects in circuits where a negative pulse is applied

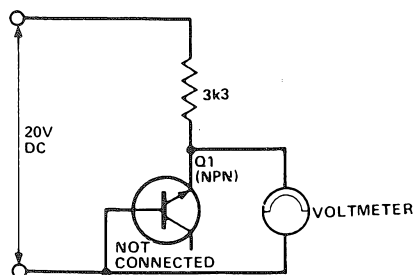


Fig. 5. Checking the zener voltages of a silicon transistor.

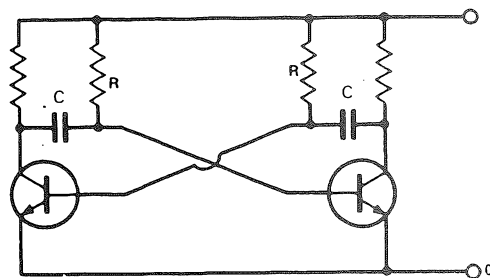


Fig. 6. Conventional multivibrator circuit.

to the base of a transistor. Multivibrator circuits, for example, operating on voltages greater than 7 V, suffer from this. Theory says that the time period of the MV is $1.4CR$ (Fig. 6), because the capacitor always charges up from $-V$ to about $0V$ whatever the value of V . The reason is that when one transistor conducts its collector voltage shoots down by about V volts, and the capacitor coupling to the next base makes that base move from about $0V$ to $-V$. Since the transistor switches on again at just above $0V$, the capacitor always charges to half way between $-V$ and $+V$, no matter what the value of V is. That theory doesn't apply if the base-emitter junction zeners, because the voltage at the base will be clipped by the zener action. We find therefore, that the frequency of the MV increases as we increase the voltage, whatever the books say about it!

Want a stable value of low voltage? Try the circuits of Fig. 7. The voltage between collector and emitter of a transistor is always low when the transistor is bottomed, with the base positive (NPN transistor) and a load resistor limiting the amount of current that can pass between collector and emitter. With the transistor the conventional way round, the voltage between collector and emitter can go as low as 0.2 V, but even lower voltages can be obtained if the transistor is inverted, with the emitter connected through the load resistor to the positive line and the collector to the negative rail. This, for example, can be very useful for clamping circuits if a small DC 'offset' is needed, but care should be taken to keep the currents low. Transistors are much more easily damaged when they are operated this way round.

Paint-scraping Saves

A few circuits specify phototransistors, which aren't always easy to obtain and sometimes (shop around!) costly. Now there isn't much you can do to make

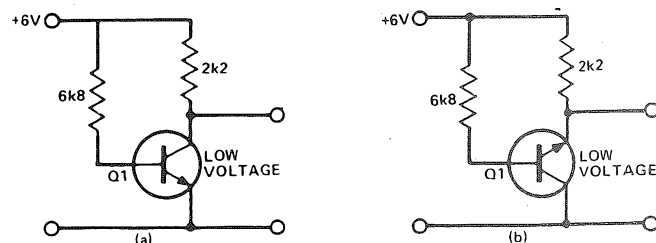


Fig. 7. Obtaining very low stabilised voltages (a) conventional method, (b) using an 'inverted' transistor for lower voltage output.

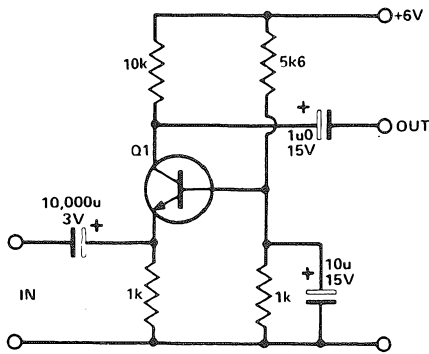


Fig. 8. Using a common-base amplifier. Note that the input capacitor must be of a very large value.

phototransistors out of modern silicon metal or plastic cased transistors, because light just doesn't pass through these materials. The old germanium transistors, like the OC72 series, were packaged in glass cases, however, and the cases then painted over. The reason for the paint is simple — any transistor junction will act as a phototransistor unless it is covered up! Scrape the paint off, and you have the phototransistor you need. Since old OC72's can often be got in lots at pennies, each, and the photo version, the OCP72, seems to fetch nearly a pound, it certainly saves money to do some paint scraping!

Ever want to drive a transistor amplifier from a really low-impedance source? There aren't many home-made ribbon microphones around, but a moving coil loud-speaker makes a useful microphone apart from its low resistance of 3R or so. Remedy here is to make use of the first type of transistor amplifying circuit that was ever used, the common-base circuit. In a common-base amplifier, the base is decoupled, with no signal input. The signal is fed into the emitter circuit, and taken in the usual way from the collector, using capacitors to keep the bias voltages correct. Advantages? There's voltage gain for a start, but the main advantage is that the input resistance is very low, offering a better match to the low resistance of the 'microphone'. Incidentally, a transistor operated this way round will amplify and oscillate at higher frequencies than is usually possible in the normal (common emitter) configuration.

Phase Splitting

This is an example of using a transistor to match impedances, like a transformer. The other impedance —

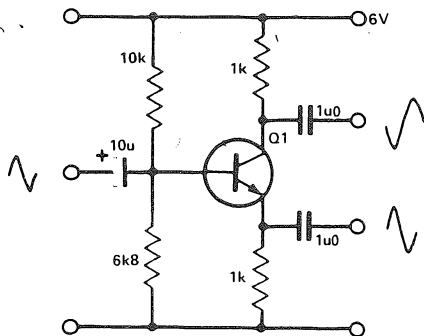


Fig. 9. The transistor phase-splitter.

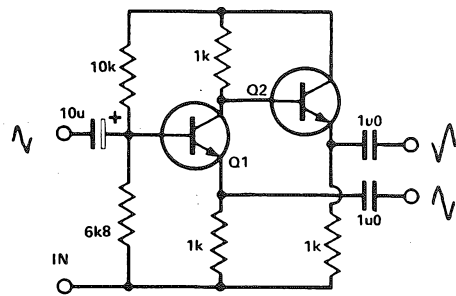


Fig. 10. Modified phase-splitter with equal output resistances.

transforming circuit is, of course, the well known emitter follower, with a high input impedance and low output impedance. If you need the phase splitter action of a transformer, but don't have a suitable transformer, don't get wound up, just try the circuit of Fig. 9. If you're driving signals into a low impedance of course, you may find that the difference between the impedance level at the collector and at the emitter causes bother (the impedance at the collector is equal to the collector load resistor, the impedance at the emitter is only a few ohms; roughly 25 ohms when the steady bias current is 1mA). In that case, another transistor added to the circuit equalises things a bit, as shown in Fig. 10.

You might think that the possibilities of the transistor were about exhausted; but we've only been using them in ones so far. When we start using transistors in twos and threes, we can substitute a lot more devices.

Unijunctions

Unijunctions, for example. Who's got a set of unijunctions around? Useful little devices. In circuits like Fig. 11 they provide an oscillator which gives a pulse output ideal for firing thyristors. The wiley experimenter doesn't worry if the unijunction drawer is empty, though. He connects up the circuit of Fig 12, which does pretty well all that a single-package unijunction will do, with the additional advantage that the firing voltage can be variable.

The action is like this. Point B, where the base of Q1 is connected to the collector of Q2 is connected to a potential divider, resistors R1 and R2. For most applications, these resistors will be equal, using (typically) 47k to 10k values. The circuit will pass no current while the voltage at point A, the emitter of Q1, is less than the voltage at point B, because Q1 is cut off (PNP,

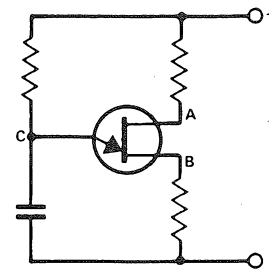


Fig. 11. A unijunction oscillator. A negative pulse is obtained at A, a positive pulse at B, and sawtooth at C.

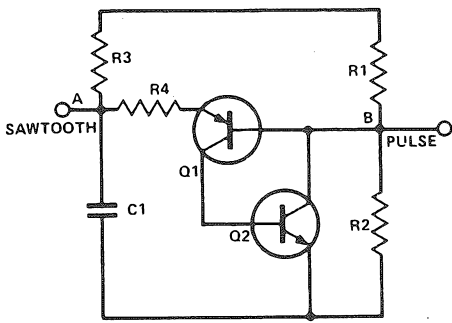


Fig. 12. A two-transistor equivalent of a unijunction.

remember), and it holds Q2 cut off as well. When point A reaches a voltage around 0.5 V higher than the voltage at point B, though, Q1 starts to conduct, and current starts to flow into the base of Q2, causing Q2 also to conduct. With Q2 conducting, the extra voltage drop across R1 causes the voltage at B to drop, dragging the voltage of point A with it. If the base current of Q1, is likely to be exceeded (as usually happens if there is a capacitor connected to point A), a small series resistor R4 (about 100R) is a good protective system. Note, by the way, that when a unijunction or this replacement is used in a timebase circuit, the value of the charging resistor, R3, must not be too low, otherwise the circuit can 'stick', not oscillating. A value of around 47k is usually regarded as a safe minimum, so that if the frequency is controlled by a variable, a 47k should be connected in series. The firing point of the unijunction substitute can be varied to some extent by making the voltage at point B variable, using a preset potentiometer, in place of R1, R2.

There is a limit, however, to the voltage range which can be used — if the voltage is too high, the circuit may not fire, if it's too low the circuit passes current continuously.

Another advantage, of course, of the circuit of Fig. 12 is that power transistors can be used. In this way, higher current pulses can be obtained than we can get from small unijunctions.

DIY Thyristor

You don't have to be stuck for lack of a thyristor, either. The circuit of Fig. 13 simulates the action of a thyristor, with the anode, cathode and gate connections as marked. With the 'gate' at cathode voltage, Q2 is shut off, so that its collector voltage is high. With the collector voltage of Q2 high, the base voltage of Q1 is also high. Since Q1 is a PNP type, having the base high means

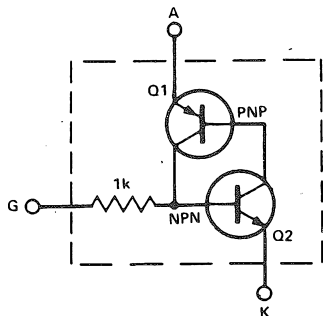


Fig. 13. Using two transistors in place of a thyristor.

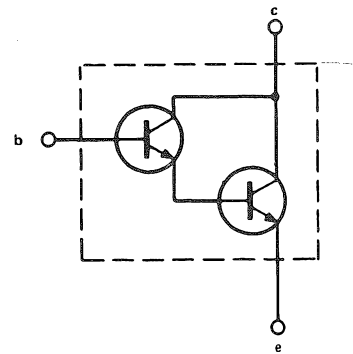


Fig. 14. The Darlington pair circuit — this behaves like one single transistor with a very high value of current gain.

keeping Q1 shut off. Now when the 'gate' lead is made more positive, so that Q2 starts to draw current, the current through the collector of Q2 is drawn through the base of Q1, ensuring that Q1 conducts. This in turn means that the base of Q2 is connected to the positive supply through the collector of Q1, keeping the pair of transistors switched on.

Don't expect to replace a large thyristor with this circuit, because the current between 'anode' and 'cathode' all passes through the base-emitter junctions. For medium-power transistors, such as the 2N2219 or BFY50 the absolute maximum base current is about 100 mA, and 50 mA is a safer limit. Power transistors such as the BD131, BD132 will stand up to 0.5 A through the base-emitter junction. The circuit will, incidentally, switch off if a negative pulse is applied to the 'gate' from a low impedance. In this respect, the circuit is similar to that of a small thyristor, most of which can also be switched off in the same way.

Changing Bias

Transistors in bunches can also be used to solve awkward problems. Suppose you want to substitute a transistor with another type which needs much more bias current. One way round, of course, is to adjust all the bias circuits. A much easier method is to make use of two transistors, with one emitter driving the base of the

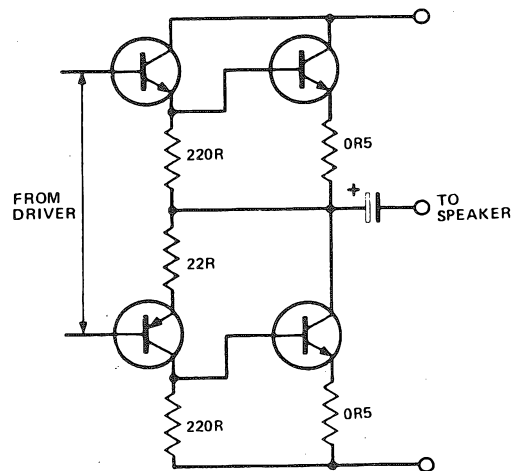


Fig. 15. A quasi-complementary output stage. The power transistors can both be NPN types.

next (Fig. 14). If the two share the same collector lead, this circuit is called the Darlington pair, but if the collector of the first transistor is returned directly to the power supply the circuit is simply an emitter follower feeding a common emitter amplifier. The difference between the two is that in the Darlington pair circuit, signal can feedback from the collector of Q2 through Q1 to the base of Q2, so reducing the voltage gain of the circuit considerably.

A two-transistor circuit can also be used to 'create' a PNP power transistor from an NPN one. The circuit uses a PNP medium power transistor (such as the 2N2905) coupled to the NPN power transistor, so that the combination behaves like a PNP power transistor. Like all two-transistor circuits, though, there is a penalty in the form of a change in DC levels. When two NPN's (or 2 PNP's) are coupled in a Darlington circuit, the voltage between the first base and the second emitter is more than 1V, when the circuit is correctly biased, instead of the 0.55 — 0.6 V we assume for a single transistor. For the PNP — NPN pair, the voltage is *less* than that for a single transistor — the base voltage of the power transistor will be 0.7 V or so above its emitter voltage, but the base voltage of the PNP transistor will be 0.6 V or so *less*, so that the DC input to the base of the PNP transistor is very close to the DC emitter voltage of the NPN one. The base-emitter voltages of these two will never be identical because the NPN power transistor will always be passing a much larger current than the PNP transistor.

Tapehead Drivers

We're still not finished with the two-transistor arrange-

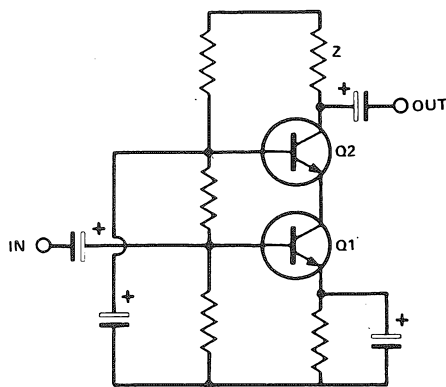
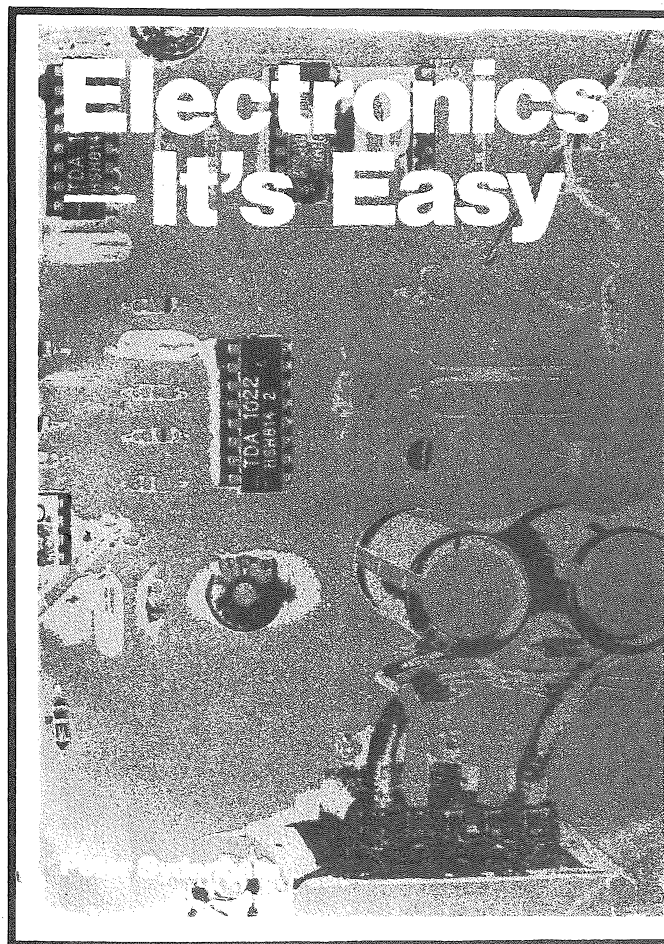


Fig. 16. A cascade stage. The load Z can be a tuned circuit or a high-value resistor providing the bias resistors are chosen to suit.

ments. Fig. 16 shows what is called a cascade circuit, with a common-emitter transistor Q1 driving a common-base stage Q2 directly coupled to it. This arrangement can also be treated as if it were one single transistor with the high gain of a common emitter transistor and the very high output resistance of a common-base transistor. It's an ideal arrangement for driving tuned circuits (because the high output resistance places very little load on the circuit) or tapeheads (because the high output resistance can ensure that the current signal into the tapehead is almost constant over a wide frequency range).

Circuits such as these described here make full use of transistors, exploiting more of their potential than the usual run of common emitter and emitter follower circuits. Make them work harder!



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DIGITAL TO ANALOGUE TECHNIQUES

Digital to Analogue conversion (DAC) is a fast growing section of electronics. Tim Orr explains some of the more practical applications.

Electronics has changed enormously in the past ten years, having swung away from valves, germanium transistors, even from discrete devices themselves. The trend is towards more and more complex integrated circuits, complete systems in a chip, large scale integration (LSI). Also the trend has swung heavily towards digitally based systems rather than analogue ones, partly because the IC manufacturers can get a greater success rate from making digital devices and partly because there are very many applications which can only be contemplated with a digital device. Such examples as pocket calculators and microprocessors spring immediately to mind. However there are several areas where analogue techniques present the only realistic solution (at this moment in time), such as tone controls in an audio amplifier. In fact, good cases can be made out for both analogue and digital systems and there are many examples where both are needed. In these it will be necessary to change from the analogue to the digital world or vice versa and to do this, some sort of conversion process has to be practised.

Digital to Analogue Conversion

The job of a digital to analogue converter (DAC) is to convert a binary code (a digital data word) into an analogue voltage. The data word is a digital representation of that analogue voltage. Thus if we presented the DAC with a digital word that was linearly increasing in magnitude, the output would be a linearly increasing analogue voltage. This digital word would be the output of a binary counter driven by a constant clock frequency. The analogue output is a linear ramp, or rather a linear staircase where the step size is controlled by the "size" of the DAC. If the DAC is an 8 bit device, ie it can accept data words 8 bits wide, then it can generate a possible 2^8

discrete output levels. Now 2^8 is 256, so therefore an 8 bit DAC could generate a staircase with 256 steps in it. The resolution of the DAC is thus 1 part in 256, or rather a change of one LSB (least significant bit) in the data word will make the output voltage change by $1/256$ th of the full scale output.

To get really fine resolution then a high performance DAC is needed. DAC prices seem to be almost linearly proportional to their resolution. I have got several DAC's amongst my collection of bits. There is an 8 bit DAC costing about £4, a 12 bit DAC costing about £35 and a 16 bit DAC costing just over £200. It is now possible to buy a monolithic (a single IC) DAC with a bit size of 6, 8, 10 and 12, but above this the devices are usually modular.

Size And Resolve

Fig 2 shows the relationship between DAC size and resolution. Notice that a 16 bit DAC with a 10 V full scale output is made up of a staggering 65,536 discrete

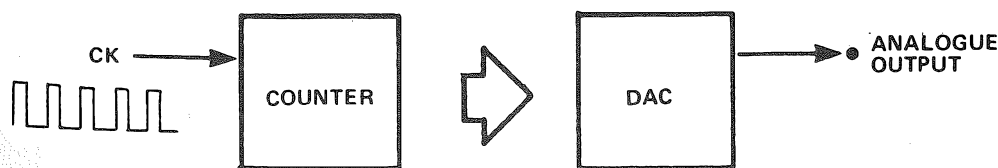
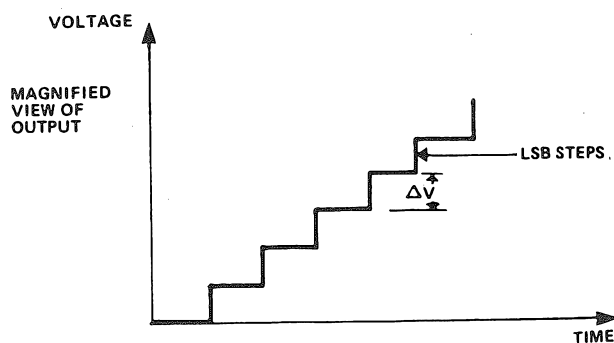


Fig 1. Converting binary code to analogue voltage.

WORDLENGTH n	RESOLUTION 1 PART IN 2^n	MAXIMUM THEORETICAL DYNAMIC RANGE	BIT SIZE ASSUMING FULL SCALE = 10V
1	2	6dB	5.0V
2	4	12dB	2.5V
3	8	18dB	1.25V
4	16	24dB	0.625V
5	32	30dB	0.312V
6	64	36dB	0.156V
7	128	42dB	78.1mV
8	256	48dB	39.1mV
9	512	54dB	19.5mV
10	1024	60dB	9.7mV
11	2048	66dB	4.8mV
12	4096	72dB	2.4mV
13	8192	78dB	1.2mV
14	16384	84dB	610uV
15	32768	90dB	305uV
16	65536	96dB	152uV

Fig 2. Relationship between size and resolution.

levels each $152 \mu\text{V}$ in size. (There is also available an 18 bit device, costing a small fortune). The larger the bit size of the DAC, the larger is the dynamic range (best signal to noise ratio) of its output. This increases by 6 dB per bit. Thus a 10 bit DAC can give a best range of 60 dB.

The human anatomy has developed over the last few million years to respond to its environment. This has resulted in the following performance figures. The sensitivity of the eye to colour is not that good. Colour television transmission doesn't give much of its bandwidth to the colour part of the signal. Have a look at a TV and see how well defined the colour is; it is usually just "sort of smeared around" the subject. Thus it is possible to get quite good digital video using only 4 bits for the colour. The eye sensitivity to resolution is somewhat better, but even so an 8 bit oscilloscope memory will look fairly continuous, giving little indication that it is made up of discrete steps.

Ear Lead

However the ear can still outperform present day technology. Using a 16 bit high quality audio system a trained ear can still detect the difference between the digitally processed sound and the original. Thus, when using DAC's in professional audio equipment great care has to be taken to eliminate all types of aberrations in the system. These digital aberrations don't just worsen the

signal to noise ratio (as an analogue system might), but they produce discordant harmonic distortion, sidebands like those obtained from ring modulation and other little funnies.

Figure 3 shows a DAC system in operation. The output of the DAC is meant to produce nice clean square wave steps, but the leading edges of these steps always have small spikes (glitches), caused by the switching times associated with the DAC's internal workings. These glitches are not regular in nature and so filtering cannot eliminate them. The glitches give the sound a "dirty" quality, or, if the system is an oscilloscope display it produces fuzzy pictures.

The glitches can be removed with a little module called a DEGLITCHER, fig 4. This is a logic controlled sample and hold which holds during the glitch period, but otherwise tracks the signal from the DAC. Thus the glitches are ignored. The output from the deglitcher then passes through a low pass filter and this removes the "stepped" quality of the signal and produces a smooth analogue output. The cut off frequency of this filter is very important and is related to the data rate of the DAC. The rule of thumb is that the filter cut off frequency should always be less than half of the data rate frequency.

Buying And Building

DAC's can be bought fairly cheaply as complete IC's or they can be constructed out of generally available parts, fig 5. This circuit uses precision buffers (a CD4041 will do), E24 resistors and a FET op amp. The buffers are run from a +10 V supply and their purpose is to provide high (+10 V) and low (0 V) outputs with low source resistance.

They are driven by a 6 bit data word, the MSB (most significant bit) thus drives the 7k5 resistor, the LSB (least significant bit) the 240k resistor. So, when the MSB changes, the output of the op amp will move by a large amount (5 V), but when the LSB changes the output will only change a little (0V156). Going from the MSB down to the LSB, each bit has only half the effect of its predecessor. This is obtained by doubling the resistor values (7k5, 15k, 30k, 60k, 120k, 240k).

A 6 bit DAC can produce 2^6 discrete output levels. Now 2^6 is 64 and so the overall resistor tolerance should be ± 1 part in 2×64 , which comes out at $\pm 0.8\%$. This type of DAC is known as a resistance ladder DAC, but in its presented form it is rather limited. For instance, a 10 bit device would require a resistor range of 1024 to 1 and a tolerance of 0.05%.

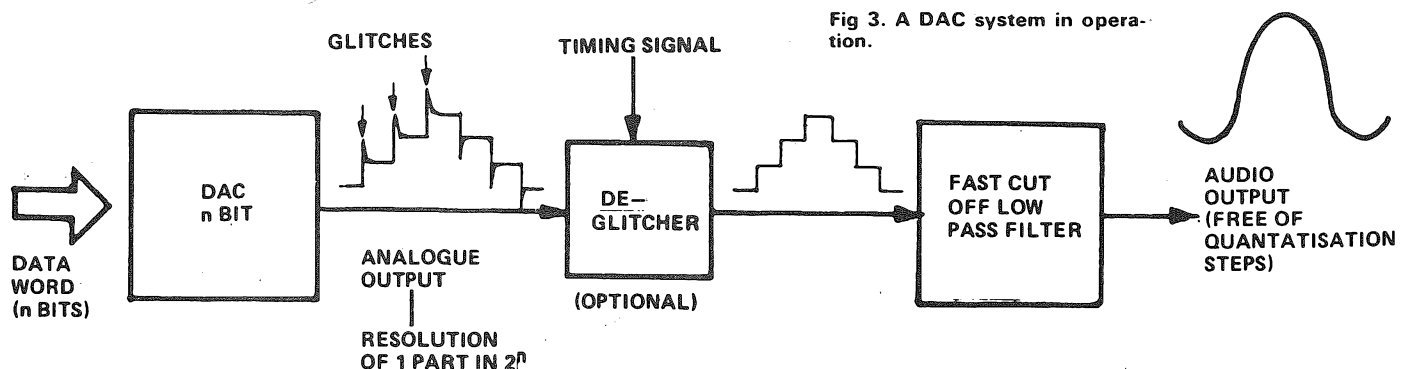


Fig 3. A DAC system in operation.

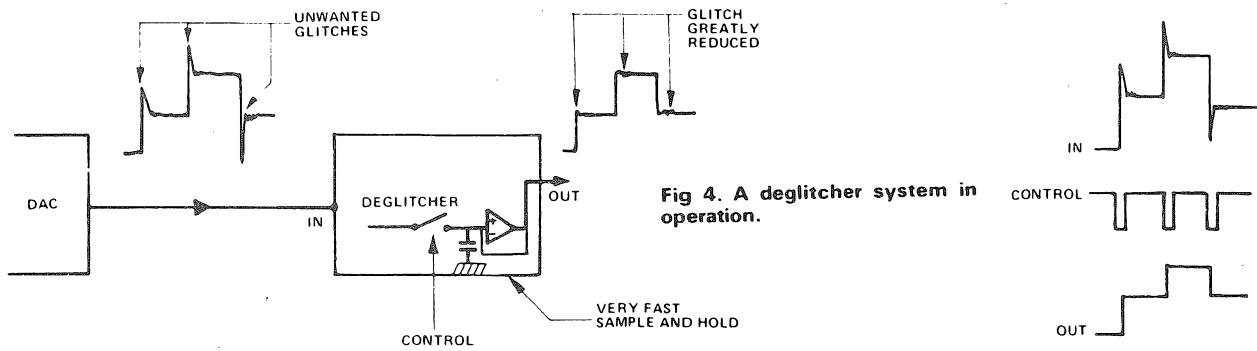


Fig 4. A deglitcher system in operation.

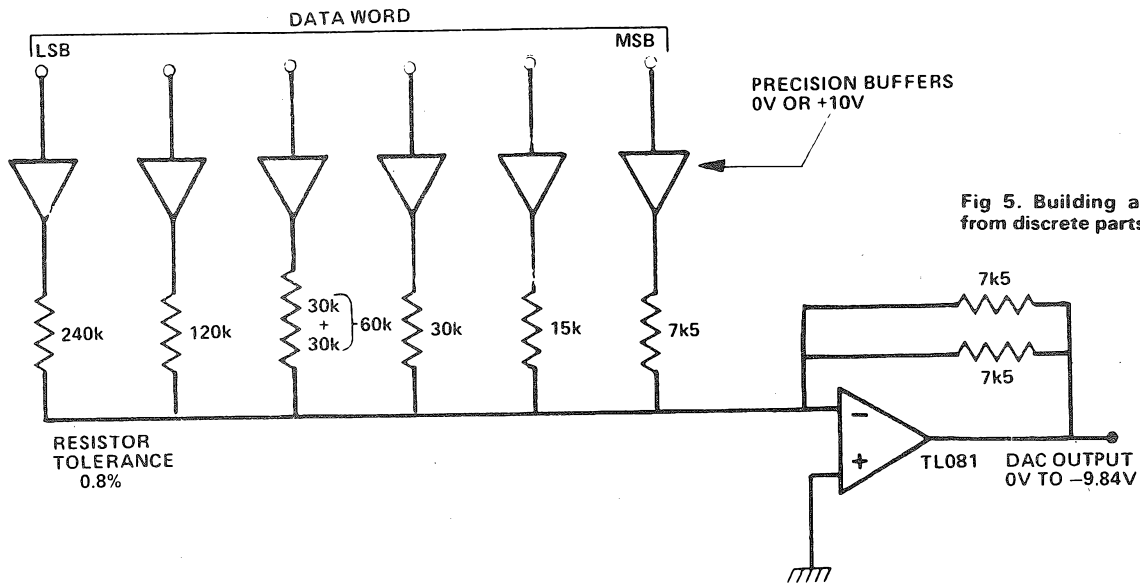


Fig 5. Building a DAC circuit from discrete parts.

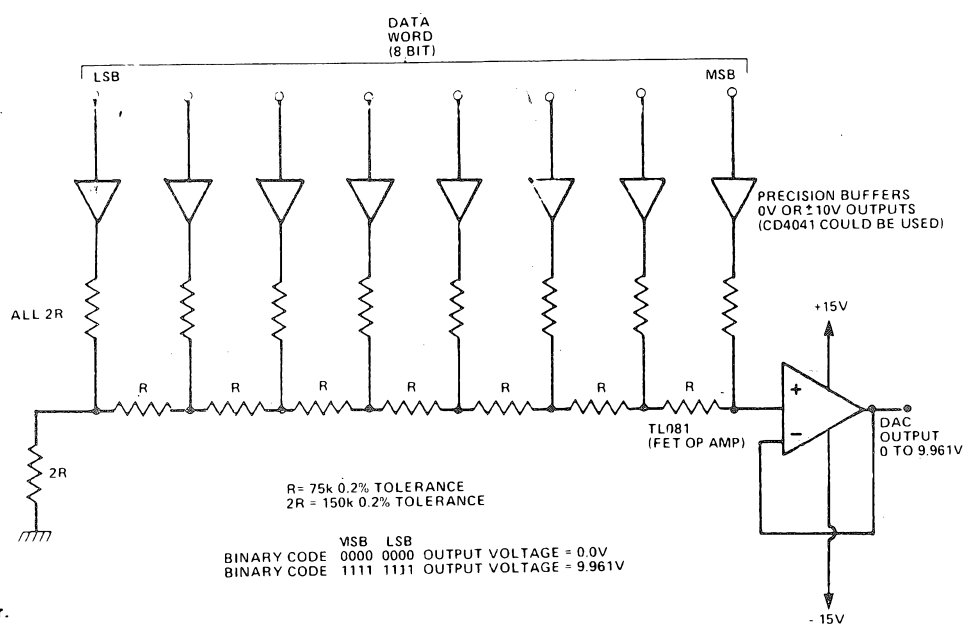


Fig 6. An R-2R ladder.

Multiple Choice

The DAC shown in Fig 6 overcomes the problem of multiplicity of resistor values; only two are needed. The resistor tolerance

still applies. Also the ratio between the resistor value and the buffer ON/OFF resistance is important. The 2R resistors connected to the buffers should ideally be 2R - (the buffer output resistance).

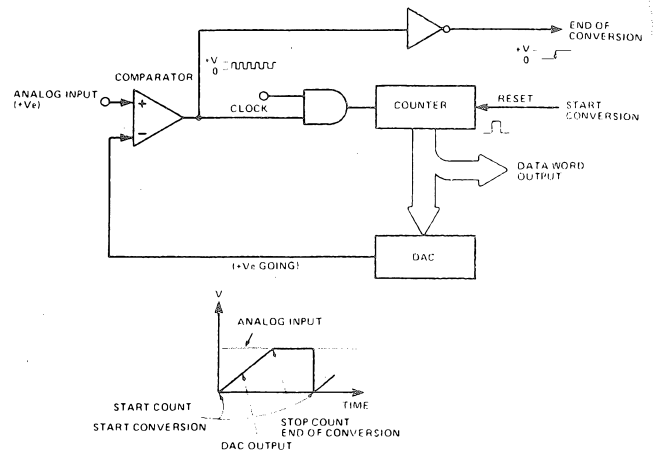
Counting On This

A "counting" type ADC is composed of a fast comparator, a gate, a counter and a DAC. This is why ADC's always cost more than DAC's, the ADC uses a DAC to do the conversion. Assuming that the analog input is positive, and the DAC produces a positive output, the conversion operation is as follows:

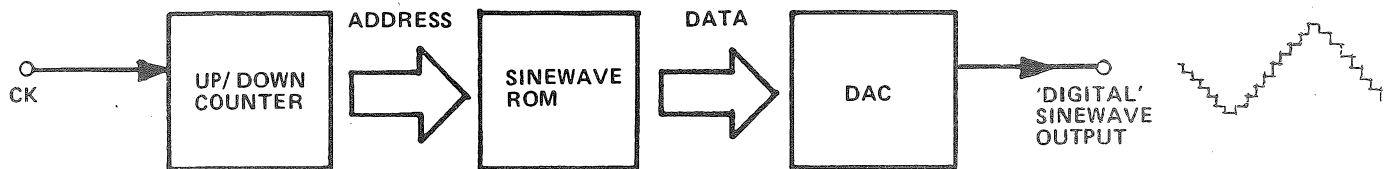
1) The signal "start conversion" is generated. This resets the counter to all zero's, the DAC output goes to zero, the comparator output goes high and so the clock is allowed to enter the counter. Thus the count proceeds and the DAC generates a positive going staircase.

2) When the DAC output exceeds the level of the analog input the comparator output goes low, the counter stops. This is the end of the conversion, and the data that is held on the counters output is the data output. It would then be transferred to some latches, and held there until the next conversion is finished.

This data word describes as precisely as is possible the magnitude of the analogue input. Although simple to operate, this method has a major disadvantage, it is slow. Imagine that the ADC is a 10 bit device and the clock frequency is 500KHz, then the longest conversion time will be 1024 counts at 2 μ Sec per count which is 2.048mSec, this means that the conversion rate will be less than 500 per second.

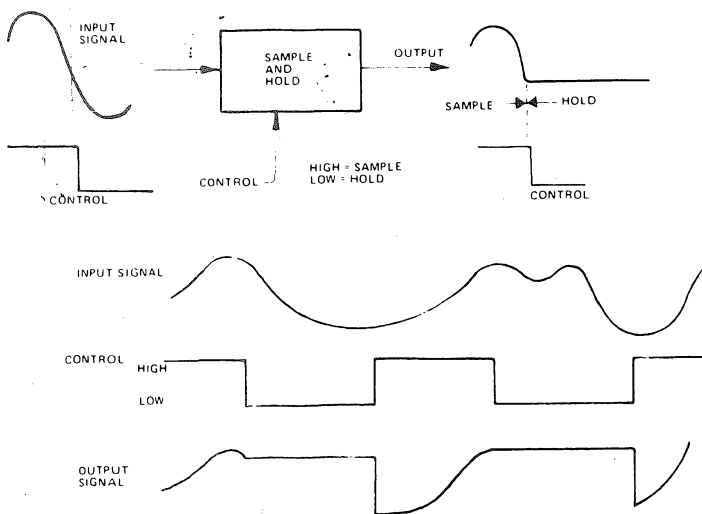


Memory Planning



The data that drives DAC's can come from several sources. It could be generated by computation or read from a programmed memory as shown. In this example a ROM (read only memory), has been programmed with the data necessary to produce a

sinewave. An updown counter provides the address for the ROM and the data is converted into an analog output by the DAC. The clock frequency divided by the size of the counter determines the sinewave frequency.



Data Lining

Another method of generating data is to convert analogue information into digital words. The signal must first be passed through a low pass filter, the cut off frequency of which must be less than half of the conversion frequency. The signal is then "held" in a sample and hold unit so that the ADC can do its conversion on a static signal. Control logic sends commands to the ADC giving it various instructions. The sequence of events is:

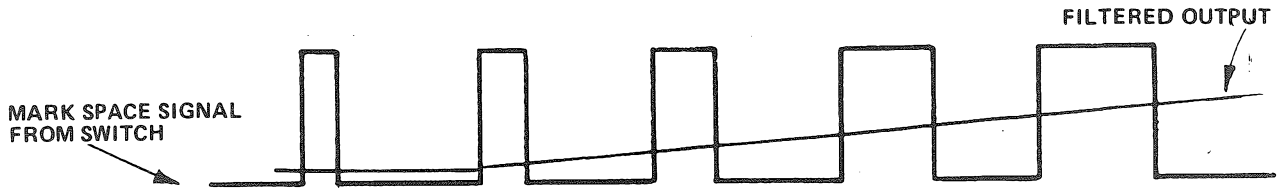
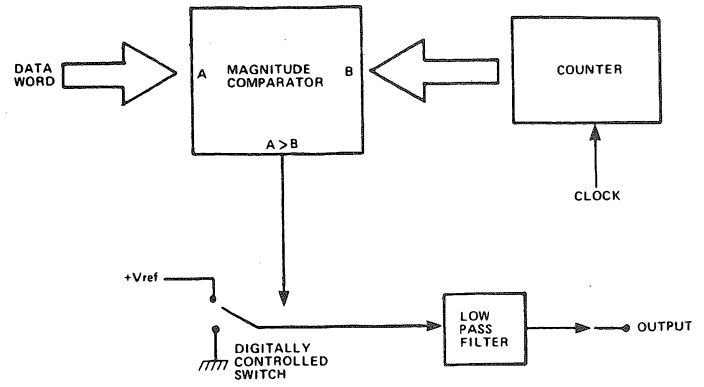
- 1) Tell sample and hold to HOLD.
- 2) Tell ADC to start conversion (SC).
- 3) Conversion finished, generate end of conversion signal (EOC).
- 4) Tell sample and hold to SAMPLE.

The process then repeats itself. The sample and hold mechanism is shown at left. Generally, in one period and the input signal several ADC conversions will be done. The data generated is then stored, processed or transmitted.

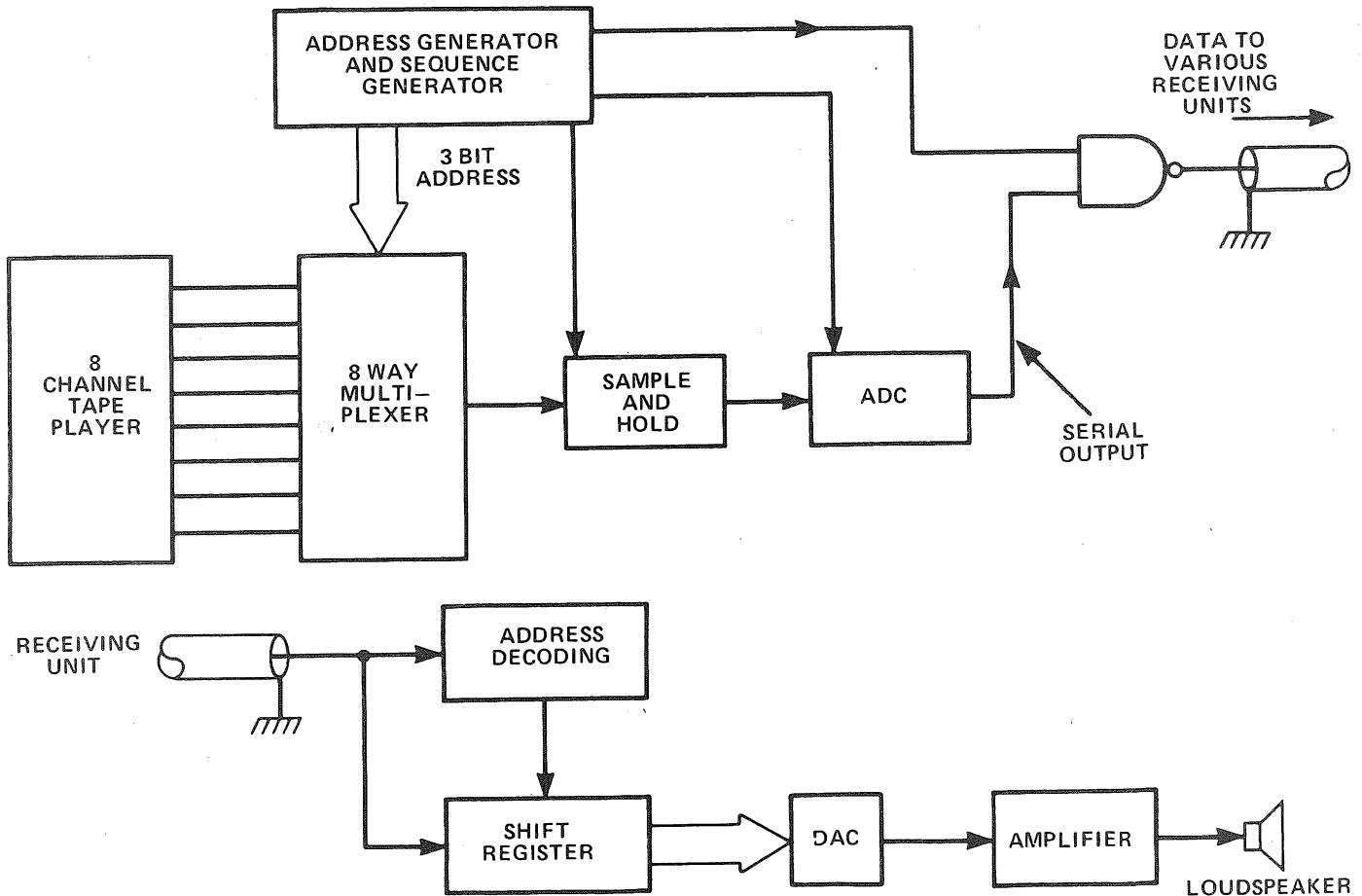


Mark Time

Yet another type of DAC, a mark space modulation DAC is shown above. The data word is presented to one side of a magnitude comparator, the output from a fast running counter to the other. When the counter is greater than the data word the $A > B$ output goes low. The output is a mark space waveform the ratio of which is linearly proportional to the magnitude of the data word. The mark space signal operates a precision switch, the output of which is lowpass filtered, providing a smoothed DC output. This type of DAC requires a fast running counter, but gives a relatively low bandwidth output signal. It is a good solution for a system where lots of slow moving outputs are required, because the counter can be common to all the DAC's.



Multiplexed Sound System



Next time you are on an aircraft with a multichannel music system, it is quite possible that the sound you are hearing via your stethoscope is digitally generated. The sounds are usually stored on a multichannel tape player and each channel is connected to a multiplexer. This is a digitally controlled rotary switch and it is continually scanning all the audio channels. The output of the multiplexer is then fed to the ADC. Thus each channel is converted to a digital code. This digital code is then

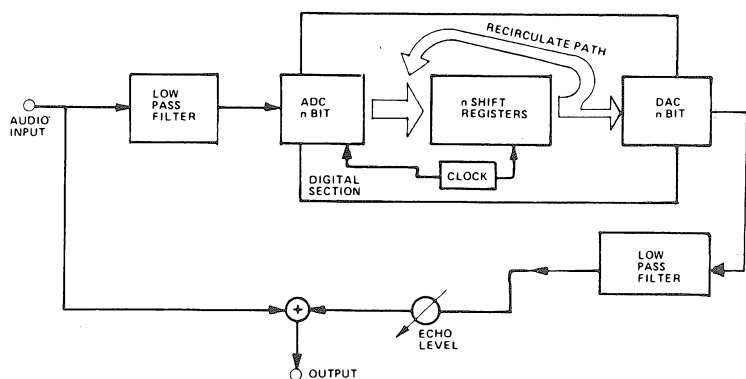
transmitted in serial mode and mixed with a sync pulse. The transmitted information is a series of serial data words, each representing a small piece of the eight music channels, plus some synchronisation data which passes down a two wire system to each receiving unit. This saves wire weight, there is less crosstalk and low pickup due to the high noise immunity of digital systems.

Digital Echo Chamber

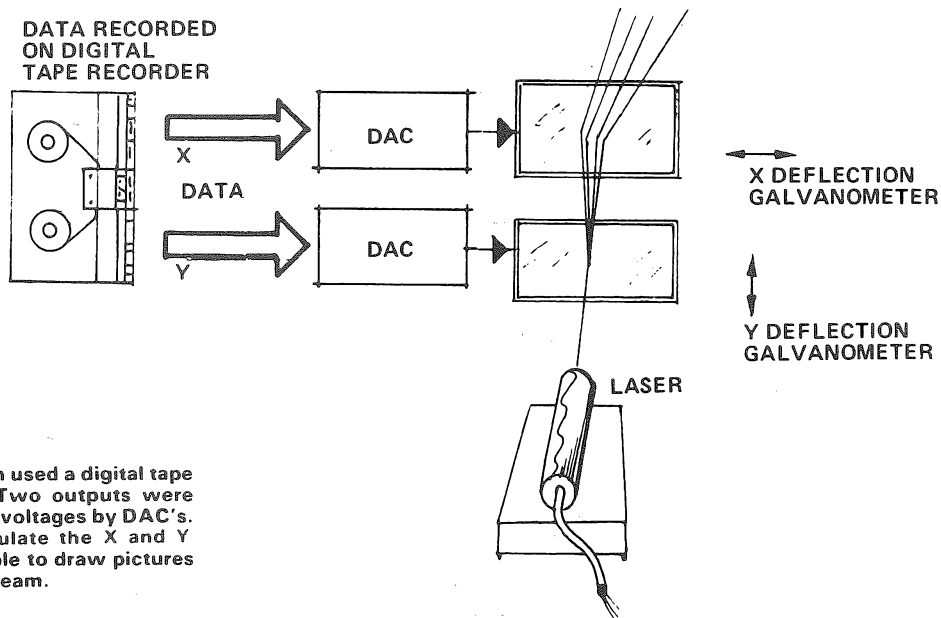
There are several professional echo chambers that are all digital. The audio input is converted into a digital word and then put into a parallel set of shift registers. A 10 bit system would use 10 sets of registers. The clock that starts the ADC conversion also shifts the data along the shift registers. The data coming out of the shift registers is then converted back into an analogue voltage by the DAC. It is then filtered and mixed with the original signal.

The echo can be made to repeat indefinitely by using the digital recirculate path around the shift registers. The amount of digital storage required is rather large. Let us assume that we want a good quality echo. This would be a 10kHz bandwidth, 60 dB dynamic range which implies a clockrate of about 25kHz and a 10 bit system. Thus to store 1 second of sound (to give one second delay), we would need $10 \times 25,000$ bits of memory, 0.25 Mbits!

The usual solution to this dilemma is to get longer delays at the expense of bandwidth. Thus a 1 second delay would be 1kHz bandwidth, a 0.1 second delay would be 10 kHz bandwidth. This would only require 25K of memory.



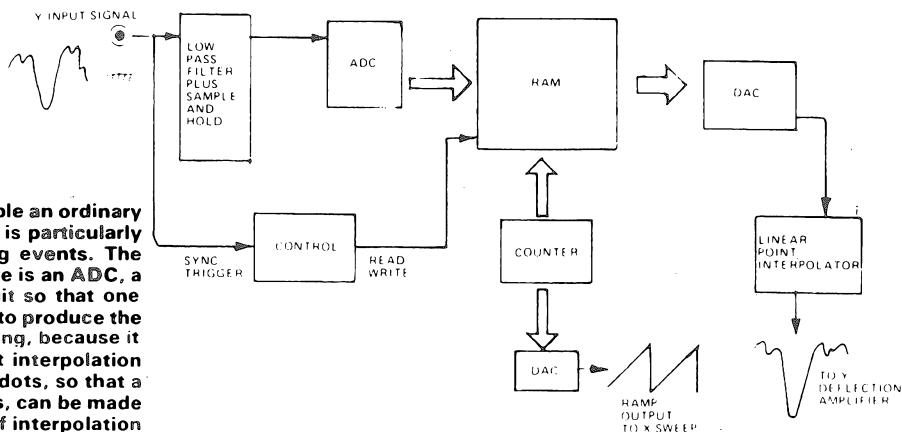
Laser Light Show.



One of the recent laser light shows in London used a digital tape recorder to store the data for the show. Two outputs were produced which were converted into control voltages by DAC's. These voltages were then used to manipulate the X and Y Co-ordinates of the laser. Thus it was possible to draw pictures and cartoon characters with a moving laser beam.

Digital Memory for an Oscilloscope.

There are several products on the market that enable an ordinary oscilloscope to store waveform information. This is particularly useful if you are trying to capture non-repeating events. The system is very similar to the digital echo unit, there is an ADC, a memory and a DAC. Also there is a trigger circuit so that one shot events can be captured and a ramp generator to produce the X sweep. The output of the DAC is rather interesting, because it is not low pass filtered, but it uses a linear point interpolation device. Basically, what this does is to join up the dots, so that a waveform that is represented by only a few points, can be made to look like the original signal. The visual results of interpolation are very good indeed.



USING THE LM3900N

THE LM3900N QUAD AMPLIFIER

One of the most noticeable trends in modern electronics is for more and more components to be packed into smaller and smaller spaces. One example of this trend is the fairly new LM3900 device which is manufactured by the National Semiconductor Company. It contains four separate internally compensated amplifiers in a single 14 pin dual-in-line encapsulation.

All four amplifiers are fabricated on a single silicon chip. Each amplifier contains seven transistors, a diode and a capacitor, whilst other internal components are used in the bias and power supplies.

One might expect that new devices of this type would be quite expensive, but the LM3900 is available at only £0.57 each in small quantities and is even cheaper in quantities of over 25 devices.

CONNECTIONS

The connections of the four separate amplifiers are shown in Fig.1. Each amplifier has a non-inverting input (marked +), an inverting input (marked -) and an output connection.

In addition, there is a single common positive supply connection and a common ground connection (negative supply line) for the whole device.

INTERNAL CIRCUIT

Conventional high gain amplifiers employ a differential input stage to provide inverting and non-inverting inputs, but a rather different approach is employed in the LM3900N. A 'current mirror' is employed in the non-inverting input circuit, the current 'reflected' in this mirror being subtracted from that which enters the inverting input.

This type of amplifier therefore acts as a differential stage by amplifying the difference between two currents

rather than the difference between two voltages (as in a conventional amplifier).

The type of amplifier used in the LM3900N may be referred to as a 'Norton' amplifier, since Norton is the name of the person who developed a theorem relating the current flowing in a circuit to the equivalent current generator and shunt impedance.

Further details of the circuitry employed in these interesting amplifiers are given in the report by T.M. Frederiksen, W.M. Howard and R.S. Sleeth entitled "The LM3900N - A New Current-Differencing Quad of \pm Input Amplifiers"; this is available from National Semiconductor (U.K.) Ltd., 301 Haspur Centre, Horne Lane, Bedford MK401TR, under the number AN-72.

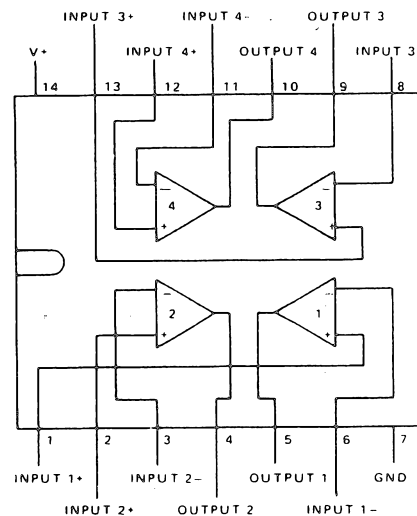
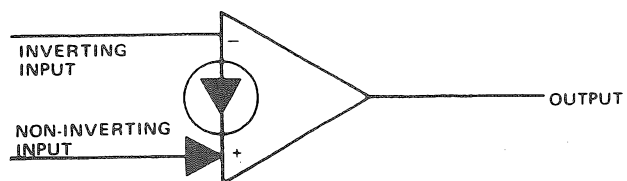


Fig. 1. The connections of the LM3900N.

Fig. 2. The symbol for one of the Norton amplifiers of the LM3900N.



SYMBOL

The symbol recommended for each of the four Norton amplifier stages in the device is shown in Fig.2. This symbol distinguishes this type of amplifier from the standard operational amplifier symbol and avoids confusion in circuits.

The symbol of Fig.2 contains an indication that there is a current source between the inverting and non-inverting inputs and implies that the amplifier uses a current mode of operation. In addition, the circuit symbol indicates that current is removed from the inverting input, whilst the arrow on the non-inverting input shows that this functions as a current input.

PERFORMANCE

The LM3900N has the advantage that it can operate from a single supply voltage over the range of 4 to 36 V. Most conventional operational amplifiers require supplies symmetrical with respect to ground (typically ± 15 V); the LM3900N can be used with such supply lines if desired.

The maximum peak to peak output amplitude of an LM3900N amplifier is only 1 V less than the supply voltage employed. The current consumed from the power supply is typically 6.2 mA (maximum 10 mA).

The typical voltage gain of each amplifier is 2800 or nearly 70 dB. The minimum gain of any amplifier is 1200. The variation of this gain with

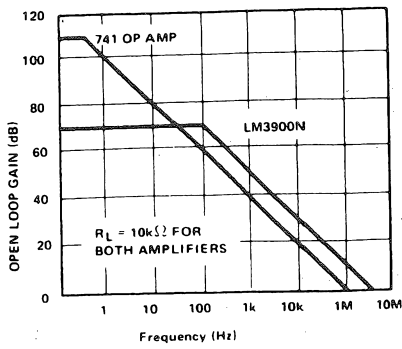


Fig. 3. Comparison of the gain of the LM3900N with that of a 741 amplifier at various frequencies.

frequency is compared with that of the well known type 741 operational amplifier in Fig.3. It can be seen that the LM3900N amplifiers provide about 10 dB more gain at all frequencies above 1 kHz.

APPLICATIONS

The Norton amplifiers used in the LM3900N device entail the use of somewhat different circuit design techniques than those used with conventional operational amplifiers.

The inverting input of the LM3900N amplifiers must be supplied with a steady biasing current. The current to the non-inverting input modulates that to the inverting input. The fact that current can pass between the input terminals leads to some unusual applications.

Both inputs of each of the amplifiers in the LM3900N are clamped by diodes so as to keep their potentials almost constant at one diode voltage drop (about 0.5 V) above the ground potential of pin 7. External input voltages must therefore be converted to input currents by placing series resistors in each input circuit.

USE AS AN AC AMPLIFIER

The LM3900N forms useful ac amplifiers, since its output can be biased to any desired steady voltage within the range of the output voltage swing. The ac gain is independent of the biasing level and the single power supply required greatly simplifies circuit design.

A simple ac amplifier circuit is shown in Fig.4. The gain is approximately equal to R_2/R_1 or 10 with the circuit values shown. The mean potential at the output is half the supply voltage. The value of R_3 should be twice that of R_2 , since the current passing through each of these resistors is then the same. The positive supply and ground connections are not shown in Fig. 4 for simplicity, but R_3 should be returned to the same positive supply line as that used to feed pin 14.

The circuit of Fig.4 provides a phase inverted output. Any ripple on the

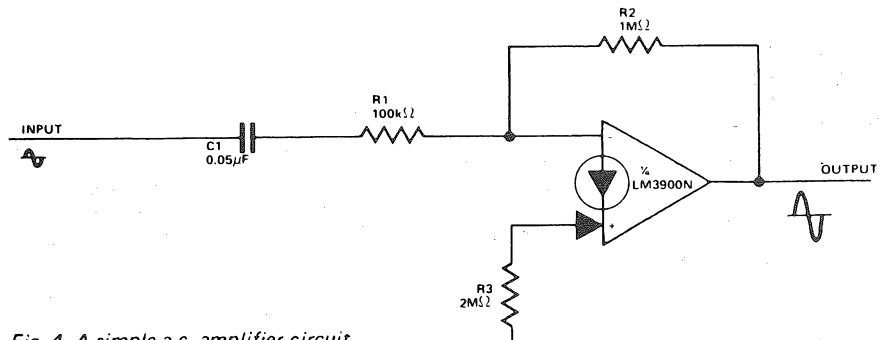


Fig. 4. A simple a.c. amplifier circuit.

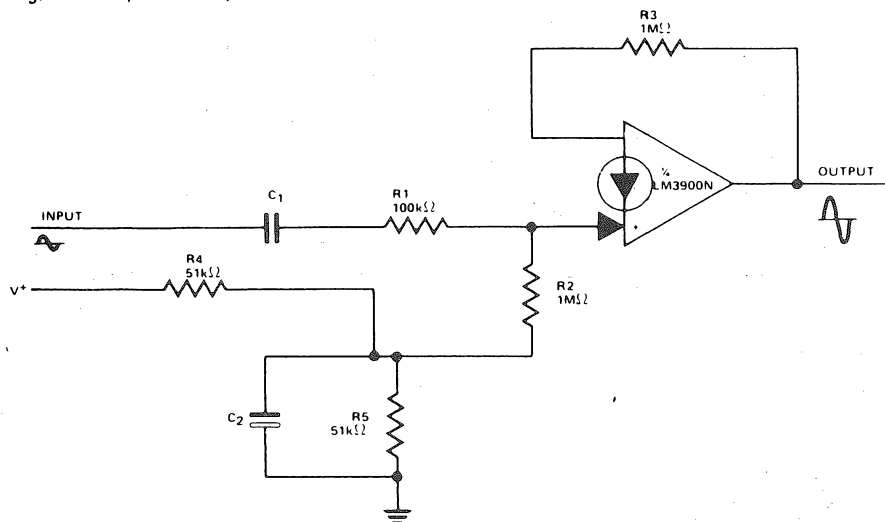


Fig. 5. A simple non-inverting a.c. amplifier.

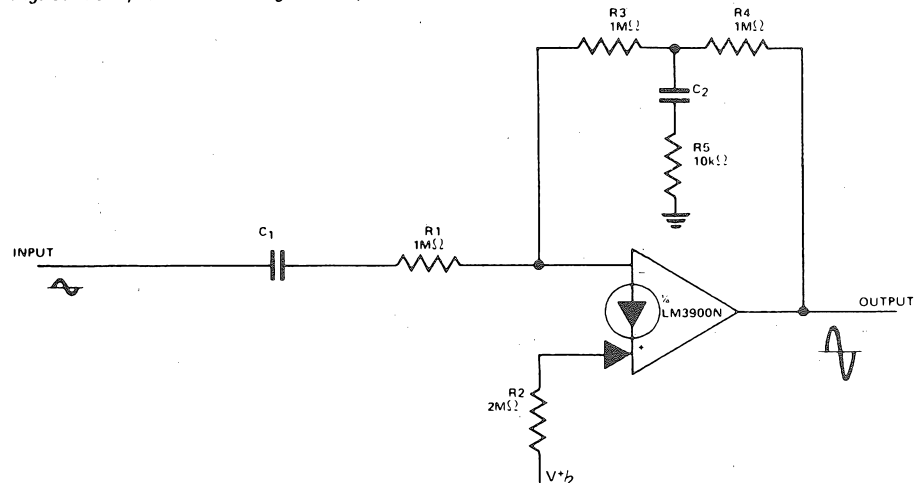


Fig. 6. An amplifier which has a high gain and a high input impedance.

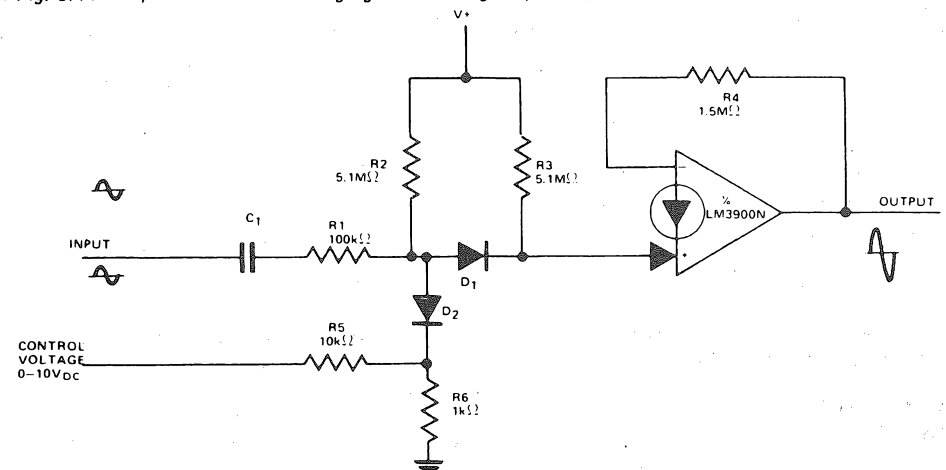


Fig. 7. An amplifier which has a gain controlled by an input voltage.

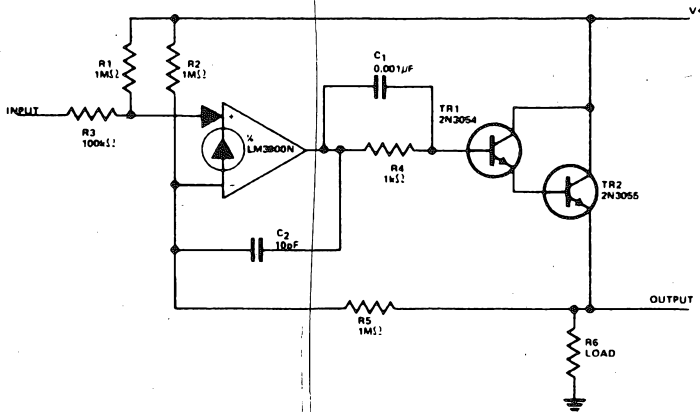


Fig. 8. A direct coupled power amplifier.

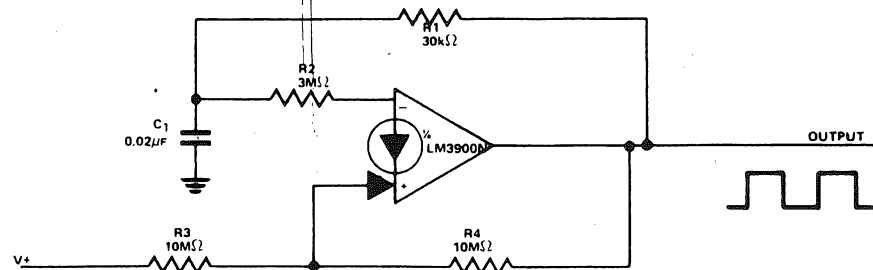


Fig. 9. A simple square-wave generator.

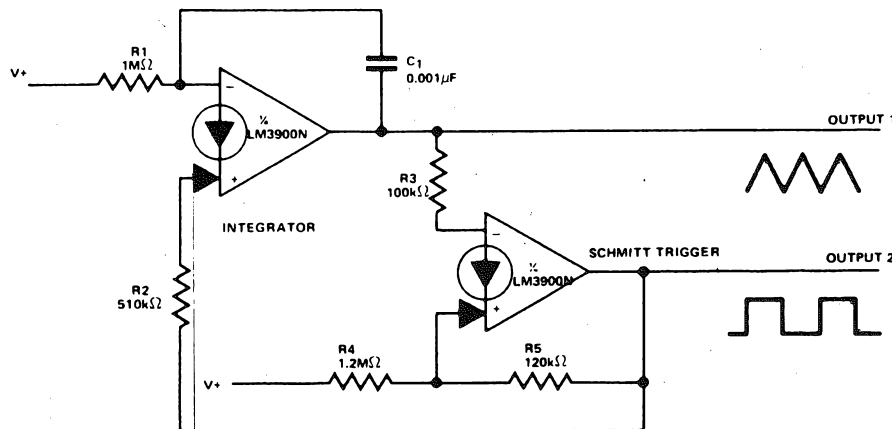


Fig. 10. A circuit for generating triangular and square-wave.

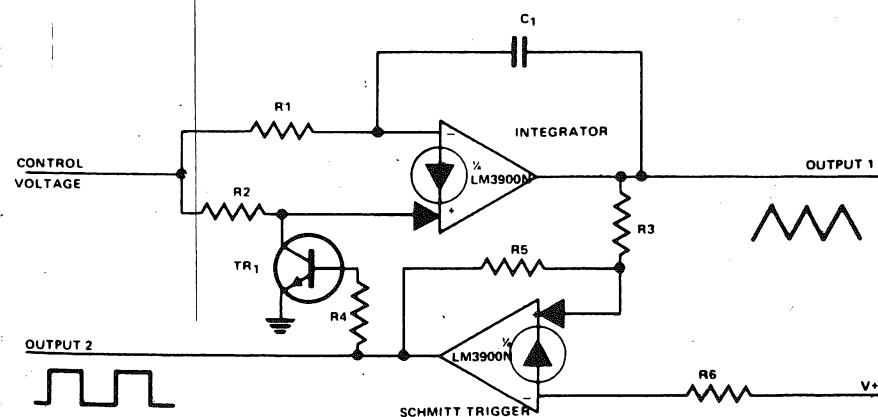


Fig. 11. A voltage controlled oscillator which produces triangular and square-waves.

power supply line will appear on the output at half amplitude.

NON-INVERTING AC AMPLIFIER

The circuit of Fig.5 shows an

amplifier which provides an output in phase with the input. The gain is equal to $R_3/(R_1 + r_d)$ where r_d is the small signal impedance of the input diode. The value of r_d is equal to 0.026 divided by the current passing through

R_2 to the non-inverting input.

The capacitor values should be chosen so that the impedance of these components is considerably less than the circuit impedance at the points concerned.

HIGH IMPEDANCE AND HIGH GAIN

The circuits of Figs.4 and 5 have an input resistance, R_1 , of 100 k ohm. If this resistor is increased to provide a higher input impedance, the gain of the circuit will fall. However, the circuit of Fig.6 has been designed so that it provides both a high input impedance and a high gain using a simple amplifier. With the component values shown, the input impedance is one megohm and the gain 100.

The voltage applied to R_2 is made equal to the output voltage (which is half the supply voltage). The value of R_2 is equal to the sum of R_3 and R_4 ; these resistors set the dc bias. If desired, R_2 may be made 4 megohms and its lower end connected to the V^+ supply.

Resistors R_4 and R_5 form a potential divider so that only 1/100 of the alternating output voltage is developed across the $C_2 - R_5$ circuit. This fraction of the output voltage is fed back to the inverting input via R_3 . As R_3 and R_1 are equal, the gain is R_4/R_5 . As R_5 is decreased, the gain approaches the open loop gain of the amplifier.

VOLTAGE CONTROLLED GAIN

An amplifier with a gain which can be controlled by the value of a steady applied voltage is shown in Fig.7.

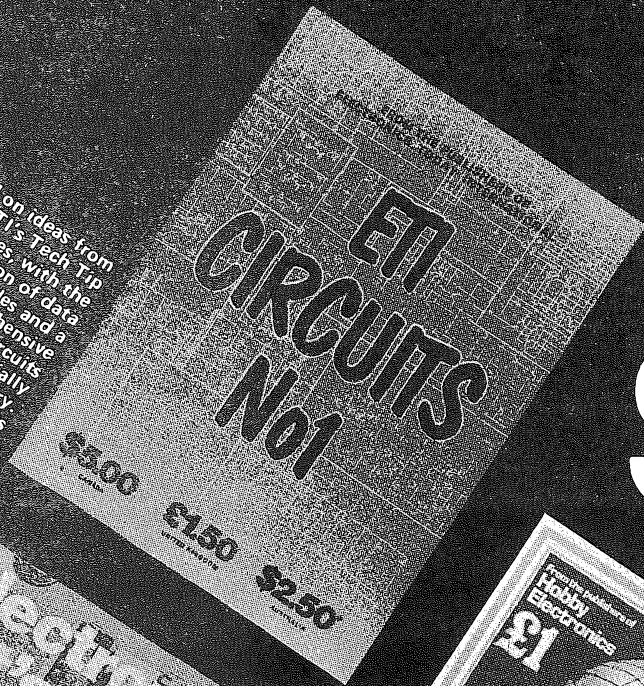
A current flows from the positive supply through R_3 to provide a bias which prevents the output of the amplifier from being driven to saturation as the control voltage is varied. When D_2 is non-conducting, the currents passing through both R_2 and R_3 enter the non-inverting input and the gain is a maximum. This occurs when the control voltage approaches 10 V.

The gain is a minimum when the control voltage is zero. In this case D_2 is conducting and only the current passing through R_3 enters the non-inverting input of the amplifier.

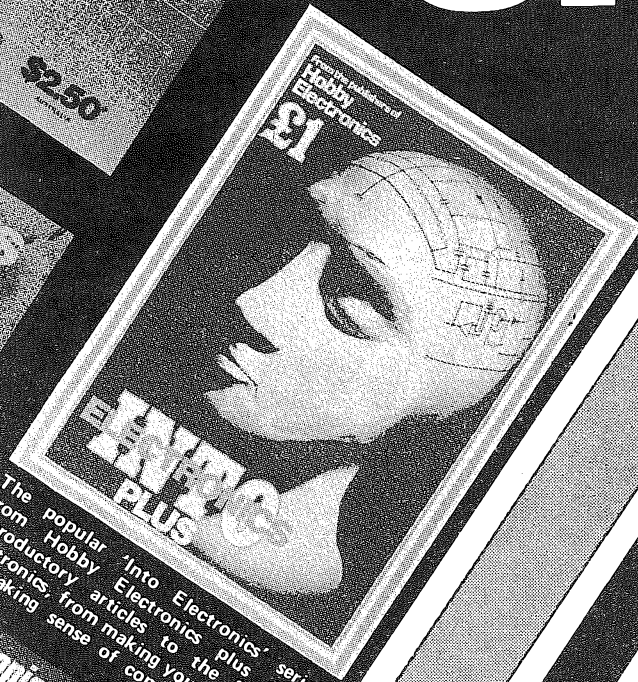
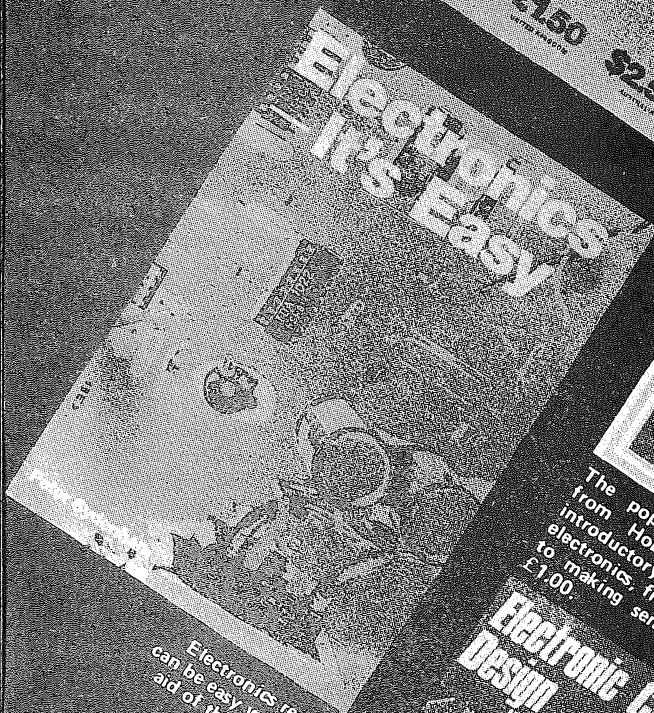
DIRECT COUPLED POWER AMPLIFIER

In the circuit of Fig.8, the output from an LM3900N amplifier is fed to a Darlington pair of power transistors. This circuit can deliver over three amps into a suitable load when the transistors are correctly mounted on heat sinks.

SPEC FR



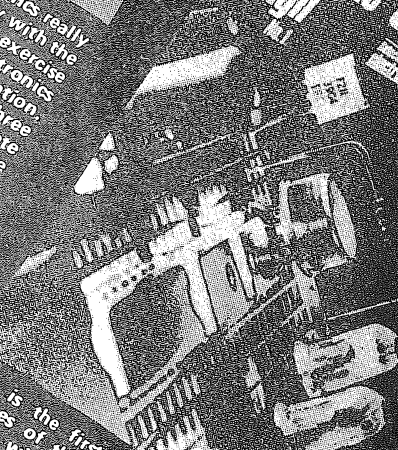
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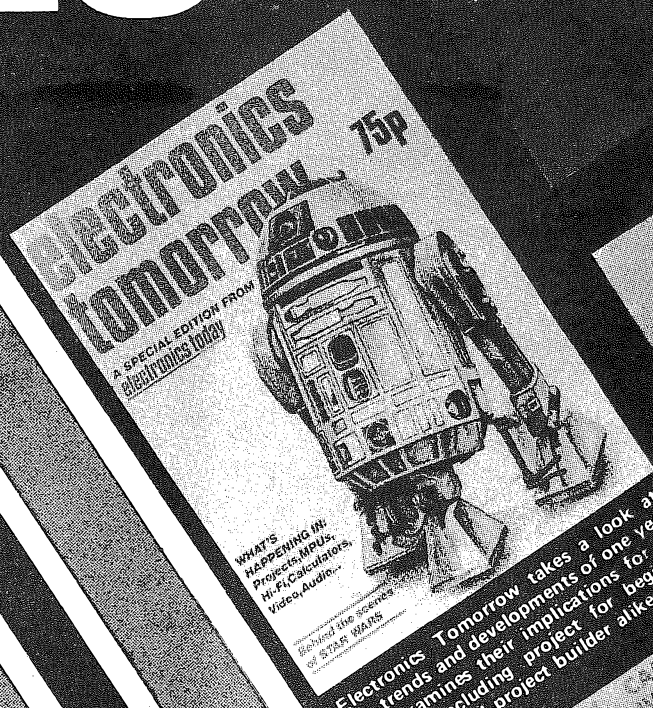
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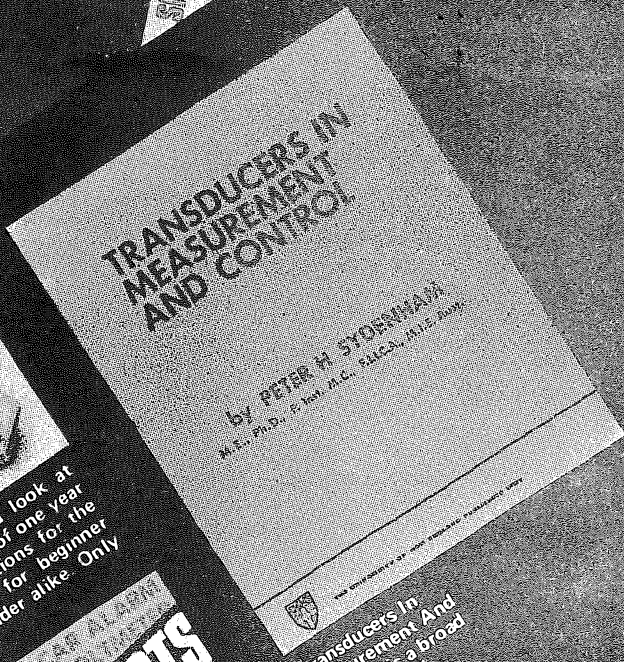
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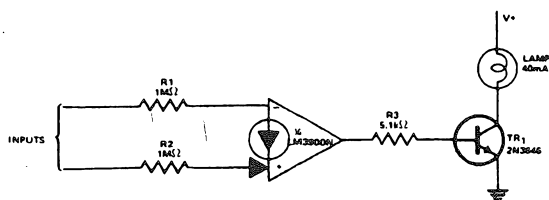


Fig. 12. A voltage comparator with an indicator lamp.

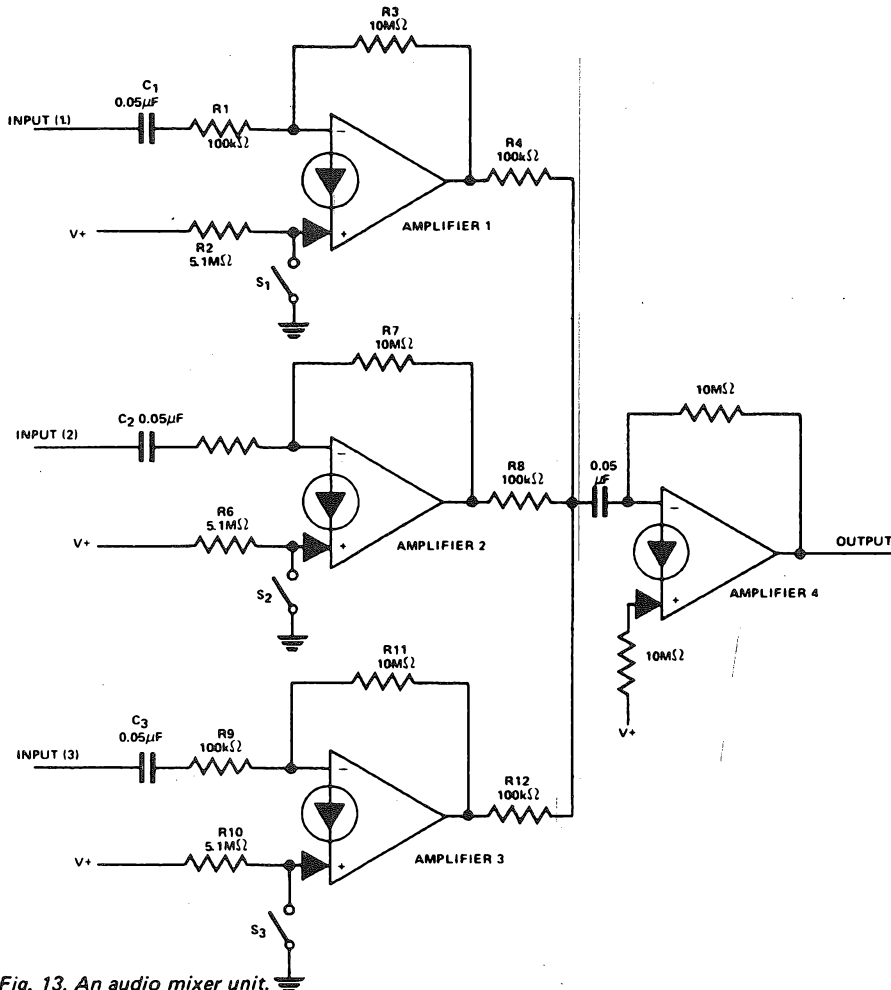


Fig. 13. An audio mixer unit.

SQUAREWAVE GENERATOR

The multiple amplifiers in the LM3900N device are very suitable for use in waveform generators at frequencies of up to about 10 kHz. Voltage controlled oscillators (the frequency of which is dependent on an input voltage) can also be designed using the device.

A simple square wave generator is shown in Fig.9. The capacitor C_1 alternately charges and discharges between voltage limits which are set by R_2 , R_3 and R_4 . The circuit is basically of the Schmitt trigger type, the voltages at which triggering occurs being approximately $V^+/3$ and $2V^+/3$.

TRIANGULAR WAVEFORM GENERATOR

A triangular waveform generator can be made by using one amplifier of a LM3900N device as an integrator and another amplifier as a Schmitt trigger circuit. A suitable circuit is shown in Fig.10; it has the unusual advantage that only the one power supply is

required.

When the output voltage from the Schmitt trigger circuit is low, the current flowing through R_2 is integrated by C_1 to produce the negative slope of the triangular wave at output 1. When the output 2 voltage from the Schmitt trigger is high, current flows through R_2 to produce the rising part of the waveform at output 1.

The output waveform will have good symmetry if $R_1 = 2R_2$. The output frequency is given by the equation:

$$f = \frac{V^+ - V_{BE}}{2R_1C_1V}$$

where $R_1 = 2R_2$, V_{BE} is the steady voltage at the inverting input (0.5 V) and V is the difference between the tripping points of the Schmitt trigger.

VOLTAGE CONTROLLED OSCILLATOR

A simple voltage controlled oscillator circuit which produces both triangular

and square wave outputs is shown in Fig.11. As in Fig.10, one amplifier is employed as an integrator.

When the output of the Schmitt trigger is high, the clamp transistor TR_1 is conducting and the input current passing through R_2 is shunted to ground. The current passing through R_1 causes a falling ramp to be formed.

When the Schmitt circuit changes state, its output switches TR_1 to the non-conducting state. The current flowing through R_2 can be made twice that flowing through R_1 ($R_2 = R_1/2$) so that the rising part of the ramp has a similar slope to the negative part.

The greater the value of the control voltage in Fig.11, the greater the frequency of oscillation. However, the voltage must exceed the constant input voltage (V_{BE}) or the circuit will fail to oscillate.

VOLTAGE COMPARATOR

The circuit of Fig.12 shows how an LM3900N amplifier may be employed to compare two input voltages and to indicate the result by means of a small lamp. If the input voltage connected to the non-inverting input is appreciably more positive than the other input, the output of the amplifier will provide a positive voltage which renders TR_2 conducting. The lamp will then be illuminated.

One of the inputs may be a reference voltage so that one can then compare a single input voltage against this constant reference.

AUDIO MIXER

The amplifiers of a LM3900N device can be conveniently used to make a mixer unit for audio purposes; the unit enables three separate audio signals to be mixed together to produce a composite output. The circuit shown in Fig.13 provides this facility using only a single LM3900N device and also enables any one channel to be selected by switches. The currents passing through the resistors R_4 , R_8 and R_{12} are summed in the input circuit of the fourth amplifier.

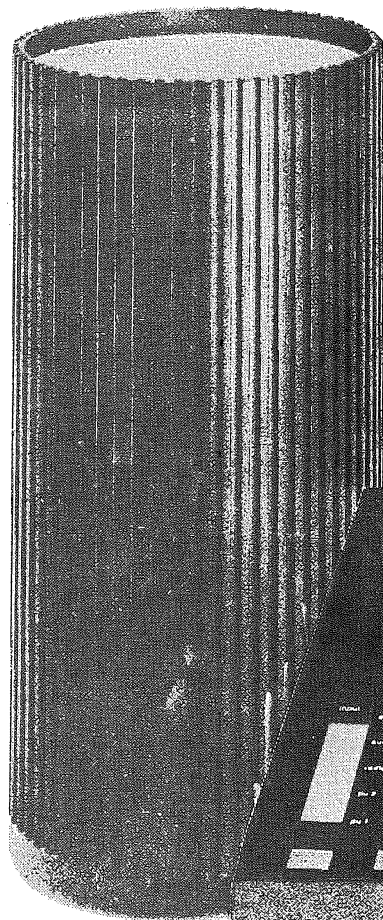
If S_1 is open, amplifier 1 will be driven to saturation by the current passing through R_2 . It will therefore be inactive.

CONCLUSION

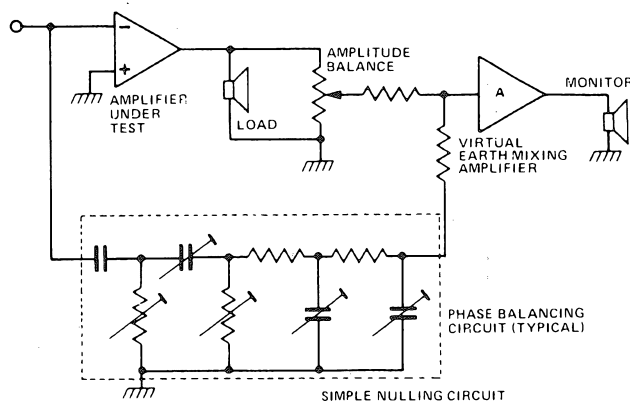
This short article has attempted to show a few of the numerous applications of this economical integrated circuit. Many more applications (such as phase locked loops, temperature sensing circuits, differentiators, tachometers, staircase generators, active filters, etc) are given in the report AN-72 mentioned previously.

DESIGNING HIGH(EST)-FI AMPS

Audio amplifier design has come a long long way since its move into semiconductors. Stan Curtis, who has been responsible for such excellent examples of the art as the Cambridge Audio and Lecson, explains here the black arts of super-fi designing.



Careful listening tests have shown that while an amplifier that measures badly is *unlikely* to sound good one that measures well *cannot* be guaranteed to sound good. This it is apparent that the traditional measurements of power distortion and frequency response need supplementing by new and more powerful laboratory tests. Such tests should more closely relate to the conditions prevailing when the amplifier is driving realistic loads and using music signals rather than sine-waves, which of course represent only one special case.



Block diagram of the Peter Walker balancing test.

Balancing Act

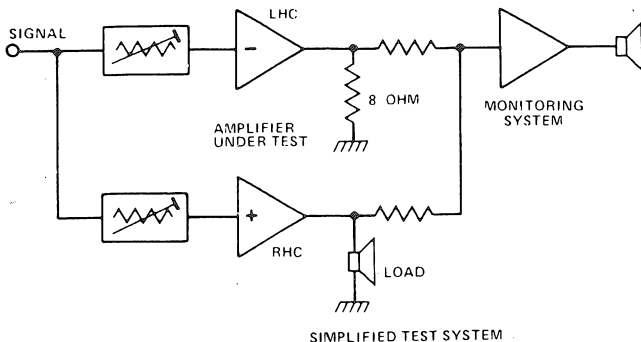
The first such test was popularised by Peter Walker of Quad. It is a simple nulling system which attempts to cancel the output and input signals of an amplifier. With full cancellation whatever remains must be distortion, i.e. signals added to or subtracted from the original. The ideal, or perfect, amplifier will produce no residual at the output of the nulling circuit and any imperfections will be monitored during a piece of music.

In practical terms the balancing of this circuit is very difficult if a significant degree of accuracy is required. Thermal drifts can aggravate the problem and generally it is difficult to set up for more than one amplifier type as usually the whole phase balance network needs to be re-calculated and re-adjusted each time. However, this simple circuit is useful for showing just how often amplifiers are clipping the signal in the course of a piece of music and how frequently some amplifiers saturate from the signal.

lecon

Sum Theory

The author used another technique at Cambridge Audio to investigate the changes in amplifier performance that are dependent upon the loudspeaker load. The two channels of a stereo amplifier are driven in mono but one channel is converted to become non-inverting. The outputs of both channels are summed and the resulting signal is monitored. Theoretically both channels should transmit the signal in the same way and (for a given circuit design) any distortion, time aberrations etc. should be the same for both channels. It is often quite possible to balance the two channels (driving 8 Ohm resistive loads) so that the residual is inaudible. However when one 8 Ohm load is replaced by a real



SIMPLIFIED TEST SYSTEM

Using one channel as an inverting amplifier to monitor distortion produced by the design.

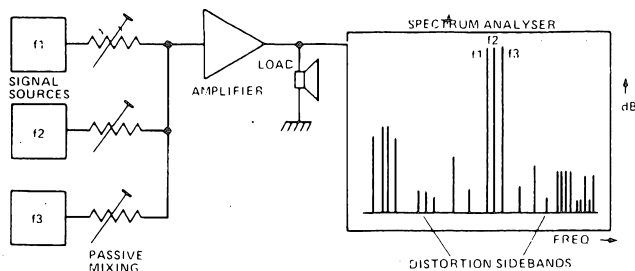
"live" loudspeaker the residual betrays problems caused by the new load. In a refined form the test works well and it did reveal two interesting things:

- i) the two channels of the average amplifier are rarely identical
- ii) some amplifiers work better in the inverting mode than in the non-inverting.

These tests serve best to indicate imperfections without generating much data to help the designer. Two simple but useful tests do generate an awful lot of usable data. The first is an HF Intermodulation Test.

IM High

The conventional IM test uses an LF (50 Hz) and an HF (7 kHz) tone in a 4 to 1 ratio and then measures the sum-total of the sideband (e.g. distortion) components. This is of little practical value unless the amplifier is particularly non-linear.



Intermodulation distortion testing using three frequencies.

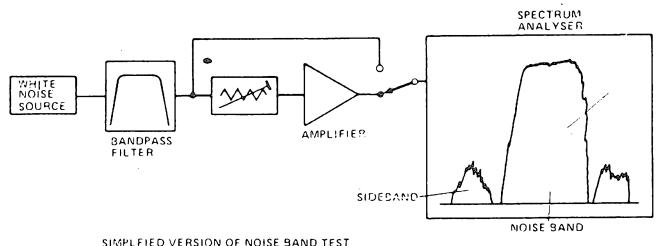
The HF IM test uses two tones of, say, 15 000 Hz and 15 100 Hz and the resulting side-bands are viewed on a spectrum analyser. The frequencies can be altered to

suit whatever simulation that is desired, e.g. two sopranos trying to sing the same note.

By repeating the tests at different levels it can be seen that many amplifiers have a performance which varies appreciably with signal level, and the test results correlate very well in identifying amplifiers with an aggressive "top end."

Dynamically Noisy

The second test is similar but attempts to measure the amplifier's performance under more varying "dynamic" conditions. A white noise source has a harmonic and amplitude structure which is variable and random and thus provides a better simulation of a music signal than does a sine-wave. The noise signal is passed through a bandpass filter to define its frequency response. The bandwidth and centre-frequency can be altered to suit the investigation as can the overall operating level. The output of the amplifier is fed to a spectrum analyser where the out of band components can be studied. Again this test is very useful for studying the effects of different loudspeaker loads but more significantly for subjecting the amplifier to random momentary "clipping" overloads.



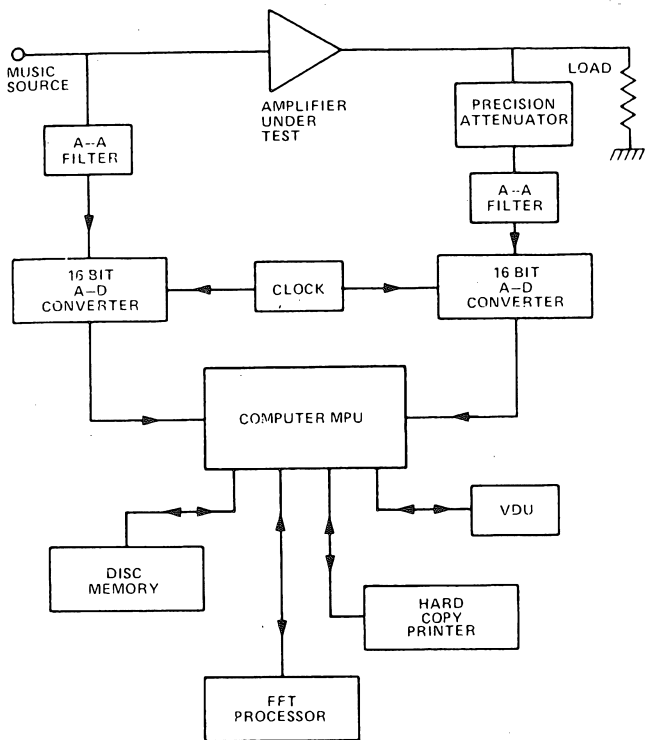
Noiseband testing with a spectrum analyser, the sidebands produced by the amp are clearly visible.

A Channel and A Log

Possibly the most complex type of testing in use is a form of input and output signal comparison used by Analog Engineering Associates in the U.S.A. and, in a simplified form, by Mission Electronics in the U.K. AEA have developed a Transient Distortion measurement system that uses music as a test signal to evaluate circuit performance under dynamic conditions. This system consists of a dual channel analogue to digital converter which is designed to have a resolution of 1 part in 65 536 or 0.0015%.

One channel of this is used to sample the input music signal whilst the second channel samples the output signal via a precision attenuator. The digitally encoded output of the converters is fed to a computer memory system for later analysis. Instead of trying to compensate for the amplifiers phase and frequency response with a passive circuit (as in the earlier simple nulling circuit) a frequency sweep is made through the amplifier to generate a "transfer function" which the computer can use to correct the data during the subsequent error analysis.

Once a series of measurements have been made in the course of playing a passage of music the resultant data can be subjected to a series of Fourier and Coherence analytical calculations. Put simply, this means that any difference between the input and output



Analog Engineering's Transient Intermodulation Distortion Measurement System, used in Britain by Mission Electronics.

signals can be described in a form that is useful to the engineer and related to the structure of the music signal at that instant. Unfortunately this test show that, as yet, no perfect amplifier exists — each type of amplifier circuit produces its own particular types of "transient error."

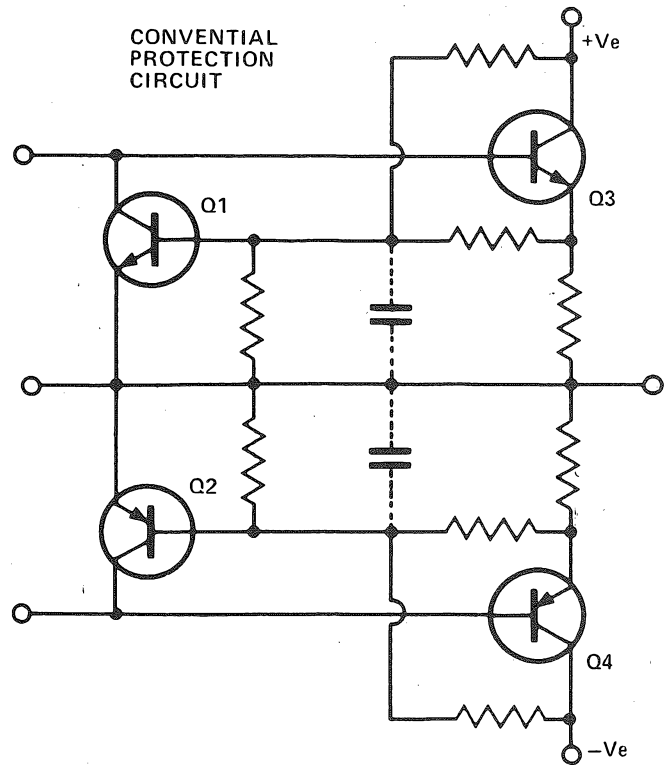
Out of The Rut

A few years ago power-amplifier design had settled into a satisfying rut. In the U.K., the Quad 303 and the Cambridge P-Series had achieved very satisfactory performance figures and they were generally considered to be good amplifiers. In the U.S.A. the Crown DC300 had achieved an almost theoretically perfect specification and was hailed as "State of the Art."

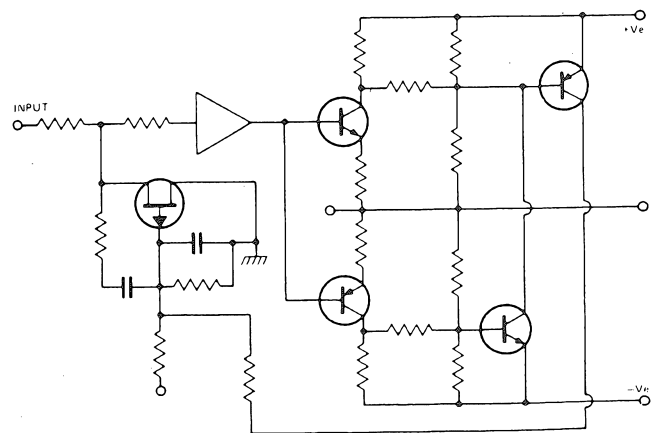
However, the first crack to appear was caused by new loudspeaker designs. Some had very demanding impedance curves which in some cases presented a 2 Ohm load to the amplifier. Such a low value of load (almost a short circuit to some minds!) operated protection circuits in many amplifiers, limiting the current to protect the output transistors.

The operation of these caused a very unpleasant "clipping" sound in some cases and even strange "clicks" and "bangs" in other cases. Thus alerted it became apparent to some designers that conventional protection circuits were turning partly-on quite frequently in the course of a piece of music and so giving a sort of premature clipping action.

Without any doubt the best results are achieved when the output stage is devoid of any protection AT ALL. The output stage should be designed to deliver all the current a load demands without limiting. Consider the reproduction of a bass drum. If the amplifier starts to limit the start of the "thump" the sound pressure will collapse and the bass-drum will appear to have no body and thus sound unrealistic.



A study of the circuit of a conventional V-I protection circuit will show that as the protection transistors turn-on they become a 'non-linear resistor' across the bases of output transistors Q3 and Q4 and as such create unpleasant distortion. One solution tried by some companies was to slug the bases of Q1 and Q2 with a capacitor to provide a time-delay to prevent the protection operating except during a sustained short-circuit.



In this protection circuit the FET starts to turn-on when full-power is delivered into a 2 Ohm load. The main advantage over a conventional protection circuit is that the limiting is "soft" (i.e. very gradual) and thus audibly acceptable and secondly that the distortion is much lower — and still only about 0.1% at limiting.

The output-stage should ideally be able to sink the full energy of the power-supply until its regulation causes the current to limit progressively. So in a good amplifier design the output-stage and the power-supply must be designed as a single item and not as separate circuits. Several amplifiers are designed like this, eg the Lecson AP3 Mk II, the BGW models 500 and 750, and the Mission Power Amplifier. The Lecson AP3/11 can, for instance, deliver nearly 20 Amps to the load before the mains fuse blows and the BGW model 750 even more.

However, with such high current capability it is essential that the amplifiers have speaker muting to prevent switch-on "thumps" (or more accurately, earthquakes) and DC offset protection to protect the loudspeakers from the effects of 20 Amps of pure DC!

Offsetting Long Tails!

DC offset has been a major problem with many DC coupled amplifiers (i.e. those having no output capacitor). The offset voltage measured across the output terminals should not be any more than ± 50 mV. Once this voltage starts to rise the loudspeaker is subjected to a DC bias which moves the coil out of the central position. This in turn causes the coil to heat up and the power-handling capability of the loudspeaker to be restricted.

Eventually (and often sooner) the loudspeaker will blow. Many amplifiers have an offset voltage that is acceptable when the amplifier is first switched on but which starts to increase as the amplifier heats up. Such amplifiers are subject to thermal drift and this drift is normally due to a component mismatch in the circuit. The conventional amplifier, with a long-tailed pair at the input, is "theoretically" free of thermal drift as these will be automatically compensated for by the DC feedback.

However, this is on the assumption that the first two transistors (or FETs), forming the long-tailed pair, are perfectly matched.

The input offset voltage (upon which the output offset voltage is dependent) is related to the base-emitter voltage V_{BE} of each transistor.

$$\text{e.g. } V_{OS} = V_{BE1} - V_{BE2}$$

This difference can be made almost insignificant by using a dual-transistor or a monolithic integrated-circuit differential stage where matching is provided by the simultaneous adjacent fabrication of the two transistors. With discrete transistors, however, a close match is unlikely.

Similarly unbalanced output loading or mismatch of the collector resistors also increases the offset voltage. These mismatches also worsen the linearity (and hence the distortion) of this stage. Thus well designed amplifiers usually use 1% tolerance resistors in these positions and adopt balanced circuitry throughout.

The offset voltage is considerably reduced by the application of local DC feedback that occurs when emitter resistors are fitted. In this case;

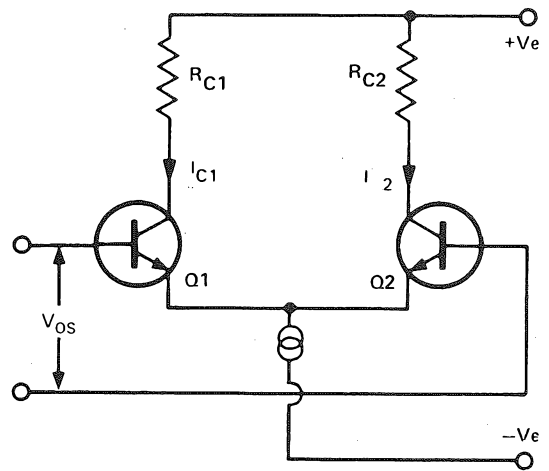
$$V_{OS} = V_{BE1} - V_{BE2} + I_{E1}R_{e1} - I_{E2}R_{e2}$$

and so by adjusting the balance between R_{e1} and R_{e2} with a trimpot a balance can be achieved.

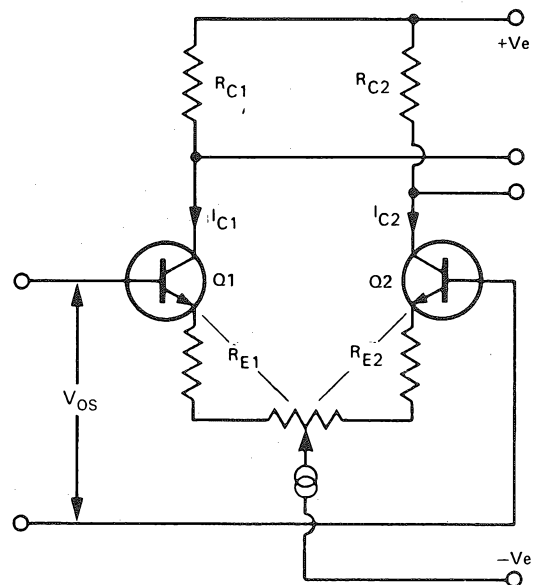
Emitting Resistance

Note that $R_e = R_E + r_e$ where R_E is the total external emitter resistance and r_e is the transistor dynamic emitter resistance. Thus it can be seen that in the earlier typical example of a stage without emitter resistors, an imbalance of r_{e1} and r_{e2} will cause a worsening of the offset voltage. More importantly it can reduce the common mode rejection of the stage. In this case the common mode is the HT lines with their ripple appearing at the output of the amplifier.

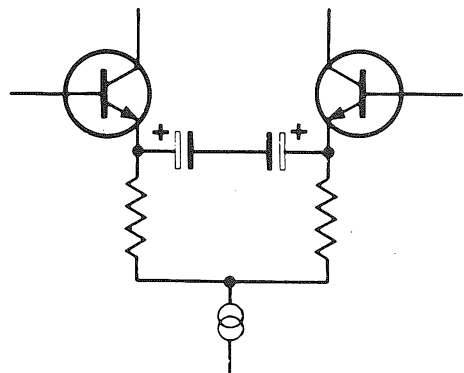
Of course the presence of emitter resistors also lowers the AC gain of the stage. For reasons to be discussed later this is not such a bad thing but in some amplifiers, for



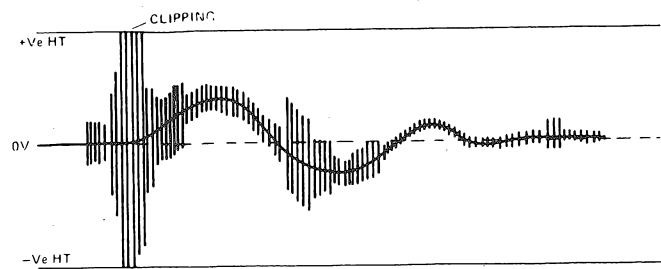
In this circuit the input offset voltage is related to the base-emitter voltage of this transistor.



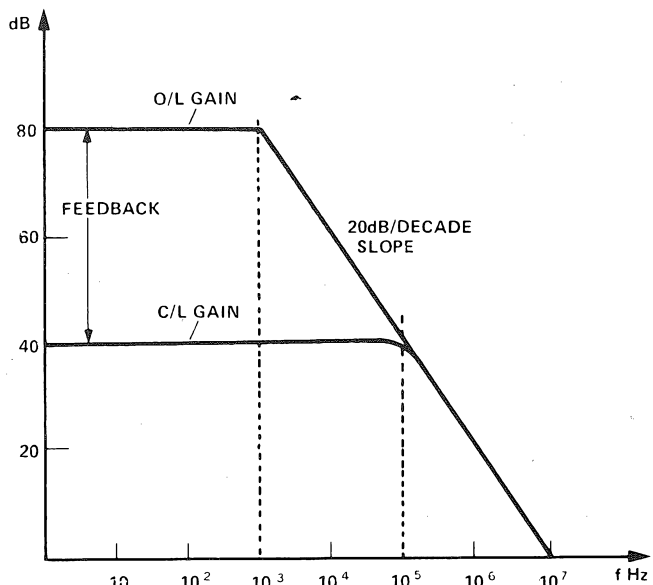
Differential pair with variable emitter resistances balanced by variation of the potentiometer.



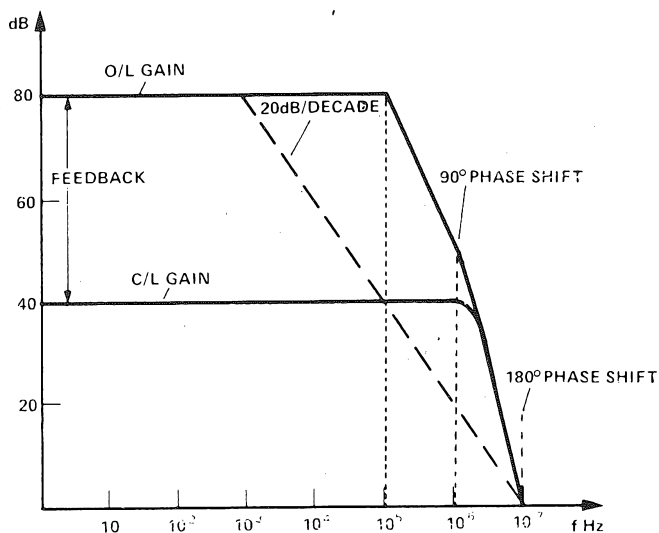
Recovering lost gain by use of bypass capacitors across the emitter resistances.



Effect of a sine wave of varying amplitude as signal upon the DC offset voltage at the output.



In the case shown in the diagram (unconditional stability) the open-loop response of the amplifier is stabilised by rolling it off at a slow 20 dB/decade slope with a single pole at 1 KHz. This amplifier would be stable with any amount of resistive feedback. However it will be seen that at higher audio frequencies the amount of feedback available reduces and so the distortion of the amplifier will increase. For this reason many amplifiers are of the "marginally stable" type.



In this case the amplifier has a fast roll-off which allows an improved closed loop performance at higher frequencies but without careful compensation they are not stable under all conditions of feedback. Once the phase shift reaches 180° the amplifier will become unstable so it can be seen that our example is only marginally stable.

example the GAS Ampzilla. This gain can be recovered by using bypass capacitors.

Clip-on Off Set

Another situation where abnormal DC offset voltages occur is following a clipping overload of the amplifier. When many amplifiers are driven into clipping, the DC voltage at the output rises towards one of the HT lines and then when the signal comes out of clipping the amplifier takes a finite time (often several seconds) to recover with the output DC voltage often oscillating between a positive and negative voltage before finally settling back to its nominal zero. Of course, when the amplifier is driven into clipping the normal negative feedback system ceases to control the amplifier.

Thus the DC instability is indicative of poor low frequency stability in the amplifier. Some of the worst (but not all) amplifiers in this respect, have separate AC and DC feedback loops and so have big electrolytic capacitors (decoupling the AC loop) which take time to charge and discharge.

The old Cambridge P100 amplifier had this problem and the effect on the reproduction of a loud bass note can be imagined to be as waffley and uncontrolled as it is. Regrettably many amplifiers still suffer from this problem.

Compensation Phase

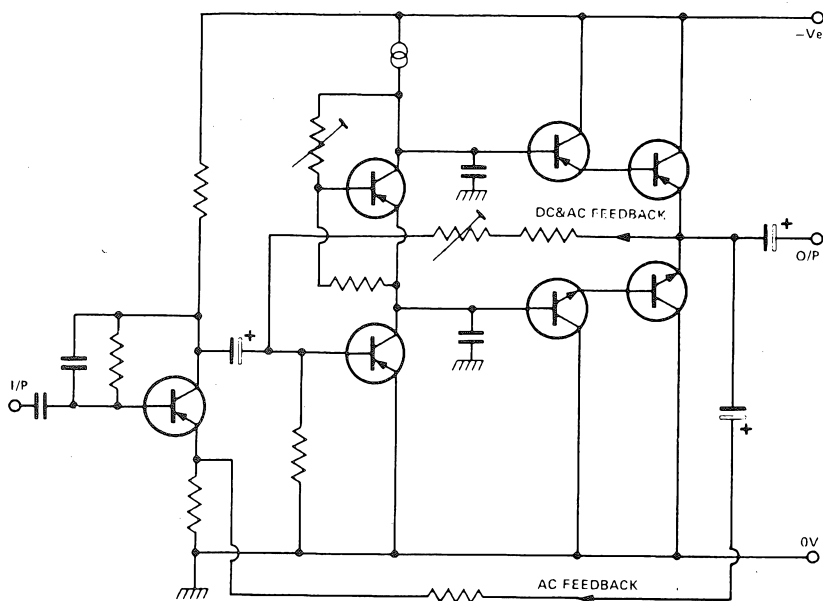
Equally important, to the owner of an expensive pair of loudspeakers, is the problem of high-frequency instability. These days very few high quality amplifiers are so unstable that they break into oscillation. However, quite a few respected units are on the edge of instability and so can potentially become unstable following a shift in operating conditions or of output loading.

Quite often some amplifiers go unstable without their owners becoming aware of the problem. Sometimes the oscillation may be moderate in level and at a very high frequency; the only symptom being that the amplifier seems to run hotter and next-door's electric drill causes more TV interference than before!

To know why some amplifiers are potentially unstable it is necessary to understand the principles of phase compensation. Many of the low distortion characteristics of amplifiers are achieved through negative feedback. If the phase shift around the feedback loop reaches 360° at any frequency at which the loop gain (i.e. the overall amplifier gain) is unity the result is a self-sustaining oscillation at that frequency.

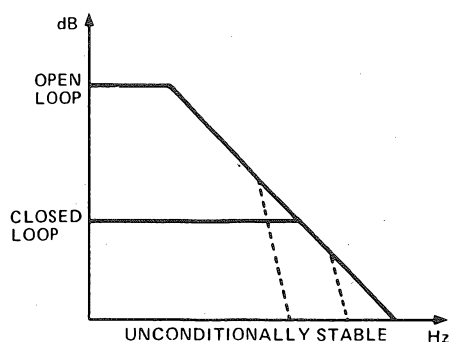
The phase-inversion to provide negative feedback produces a stabilizing 180° (ie "out of phase") phase shift, but an additional 180° can be developed in the amplifier.

The phase shift developed through an amplifier is the combined phase shift of its several stages, and it usually develops 180° at higher frequencies. To ensure frequency stability under feedback conditions, phase compensation *reduces* the amplifier gain at those frequencies for which phase shift is high and it reduces high frequency phase shift by accepting a greater phase shift at low frequencies. This is accomplished by adding response poles and zeros in the form of resistor-capacitor networks (real or inherent in the transistors) in the amplifier circuitry.

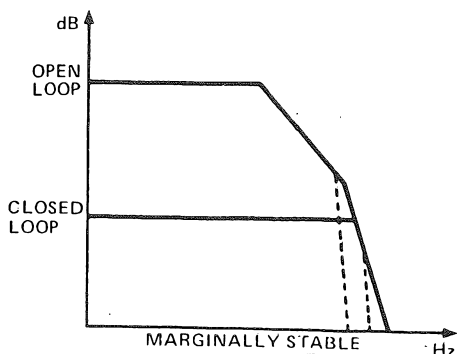


Circuit diagram showing a typical circuit which would prove to be prone to DC instability when in use. Note that separate paths exist for AC and DC feedback.

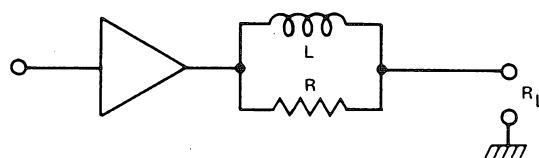
If the amplifier now has to drive a capacitive load eg. electrostatic speakers, or complex crossover networks; another pole is added at the output eg:—



Above: Effect of adding an extra pole at the output of an unconditionally stable amplifier, such as might be added by a complex crossover network. Below: Same condition applied to marginally stable type. Phase shift now borders on 180°, i.e. oscillation.



In the case of the unconditionally stable amplifier the only ill-effect will be some "ringing" in the closed loop step response — but in the case of the marginally stable amplifier it may go completely unstable. The most popular "belt and braces" solution to this problem is to fit a resistor-inductor network at the output to "cancel-out" the effect of the capacitive loading, thus:



Ever wondered what this circuit in the output of an amplifier is for? Wonder no more — it's to aid the output stage in handling a capacitive loading by partially cancelling the effect.

It is interesting to note that some marginally stable amplifiers omit those components as, in practise, most speaker cables have sufficient resistance and inductance. However, some of the new "Super-Cables" (Litz and Lucas, etc) have a very low resistance and almost no inductance but some capacitance — and their use with certain amplifiers has caused instability, with the amplifier (or speakers) eventually blowing-up!

DESIGNING HIGH (EST)-FI AMPS

PART 2

Stan Curtis considers which parameters matter most in super-fi audio, and how they can best be optimised.

For many years it has been the standard practise to specify and compare amplifiers through their ability to handle a continuous (steady state) sine-wave signal. Thus such a signal is used to measure power-output, frequency response, harmonic distortion, crosstalk, input overload capability, intermodulation distortion, damping factor, and gain! Unfortunately many engineers and Hi Fi pundits still believe that such information is ALL that is necessary to quantify an amplifier's performance and to compare it with others. Not so.

Steady-state sine-wave testing can however, tell only part of the story and can often be misleading. Music contains complex wave forms with a spectral content of greater than eight octaves and dynamic ranges of up to 100 dB. Yet such complexity is readily understood by the human brain which, in mastering the subtleties of spoken language, has evolved the ability of extraordinary auditory sensory perception. The music signal, as with all audio signals, can be considered in terms of two variable qualities — the frequency domain, and the time domain.

Frequently Timely

The frequency domain is the area that has monopolised engineers thought for so long. Even the most complex music signal can be represented by a Fourier Analysis.

This develops a mathematical equation which lists separately each frequency, making up the signal, together with its phase and amplitude. However, a Fourier Analysis is only complete in the case of simple waveforms, with more complex waveforms it becomes only a convenient approximation.

Of course, in order to make a Fourier analysis of a signal the components of that signal have to be analysed over a period of time such that complete cycles of the lowest frequency can occur.

Thus we take consideration of the Time Domain.

Where steady-state signals are concerned the Time Domain is not normally considered as the signal is of a continuous unchanging nature between any two periods. If the "time window", during which the signal is Fourier analysed is reduced progressively it becomes apparent that an accurate spectral analysis becomes less possible. It can then be seen that the important characteristics of the signal are amplitude and rate of change. In other words it's envelope.

What is required is the amplification of an audio waveform in such a way that the ear can detect no degradation.

What Do We Want?

Let us consider ways in which such degradation can occur. The waveform envelope can be distorted by amplitude changes of any component or by changes in

the phase relationship of the component harmonics.

Experimental work has established that changes in the relative amplitudes of the harmonic structure of the waveform are readily detectable.

Other work has shown that the qualitative characteristics of a complex sound depend upon the phase relationships of the component harmonics. It would seem that as a phase difference must be interpreted as a time delay between the component parts of the signal then a sufficient phase shift in a system must eventually become audible as these component parts are moved in respect to each other in time. In practise large phase shifts are very audible and indeed telephone lines are often subjected to phase and delay correction to render speech intelligible. However, establishing an acceptable degree of phase shift is extremely difficult.

Following the arrival of the "linear phase" loudspeakers great controversy has raged over whether phase shifts effect sound quality. A study of the experimental work performed to date shows that

i). It seems to be very difficult to repeat someone else's experiment (and get the *same* results!)

ii). It seems, on balance, that where recurrent waveforms (steady state) such as sine-waves (and instruments producing a "continuous" although decaying tone) are concerned; then quite large phase shifts, between the extremes of the frequency band, have no identifiable effect on sound quality.

However, a phase non-linearity on the leading edge of a true transient appears to be audibly more perceptible, particularly on speech and percussive sounds.

Bandwidth and TID

Transient signals cause many problems for amplifiers of which phase linearity is but one. Other problems are; instability and ringing, clipping, slew-rate limiting, and transient intermodulation distortion. Transient intermodulation distortion (TID or TIM) is an effect that has been much in vogue in the past 3 or 4 years but which is often misunderstood. TID can be predicted mathematically but such a description is out of place here. TID most commonly occurs when an amplifier, with overall negative feedback over several stages, is driven by a large enough signal whose frequency (or equivalent rise time) is above the open loop bandwidth of that amplifier.

Because the feedback loop is fed from the output of the amplifier, it cannot be operating until signal current flows at the output. i.e. during the open-loop rise time of the amplifier.

The outcome is very large signals occurring in the intermediate stages of the amplifier causing those stages to distort or even to clip. With some amplifiers this clipping (which cannot occur with any steady-state signal) can cause the stage to latch-up for a time until the

operating conditions restabilise.

Thus not only is the leading edge of the signal severely distorted — in some cases it is removed completely.

TID is therefore a form of overloading that is dependent upon both amplitude and time. This is audibly (but at a higher signal level) similar to cross-over distortion, as both effects cause phase and amplitude modulation of the signal due to momentary change in gain. (Remember that at the cross-over point zero, there is no current flow in the output stage and hence no feedback current and so the amplifier is momentarily open-loop).

Making Big Bands

TID can be avoided by careful design of an amplifier whose open-loop bandwidth is greater than the highest frequency of the input signal. The maximum bandwidth can then be defined at the input by a passive RC Filter. Thus if we decide upon a maximum signal bandwidth of 20 KHz then our filter will limit the signal waveform rise-time to

$$T = \frac{0.35}{20\text{kHz}}$$

i.e. 17.5 uS

Our amplifier's open-loop bandwidth should be designed to be, say 23 kHz, giving it an open-loop rise-time of 15 uS. and freedom from TID. If however, in the interests of a good specification, and possibly better reproduction, we decide upon a closed-loop bandwidth of 100 KHz (i.e. a rise time of 3.5 uS.) then our amplifier will need an open-loop bandwidth of greater than 100 kHz to maintain freedom from TID effects. In a power amplifier such performance is not easy to obtain.

Fast power transistors are notoriously easy to blow-up and are expensive. The common form of lag compensation (used where the open-loop bandwidth is perhaps 2 kHz) has to be replaced by lead compensation.

Another technique is an extension of the first in that the preceding stage of the power-amplifier is designed to have a lower open-loop band width than the next.

Important or not?

Many people now consider that TID is unimportant or that it doesn't exist. This is partly because it is very difficult to measure and only readily visible in the laboratory in the "clipping" stage. To reach this stage with most amplifiers (but not TID — free designs) requires either fast rise-time or high signal levels or both; conditions that are unlikely to occur in practise.

However, a large degree of non-linearity and hence bad intermodulation will still occur with more realisable input signals. Although this cannot be measured yet (how do you measure say, 5% IM over a period of 5 milliseconds??) it can be predicted mathematically and, just as important, heard. Amplifiers free of TID have a very "open" quality with accuracy of depth.

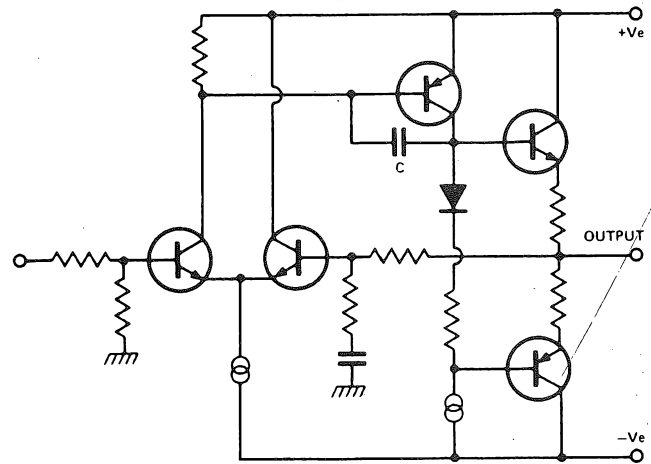


Fig 1. Circuit diagram of a typical amplifier circuit which employs lag compensation techniques — provided by C.

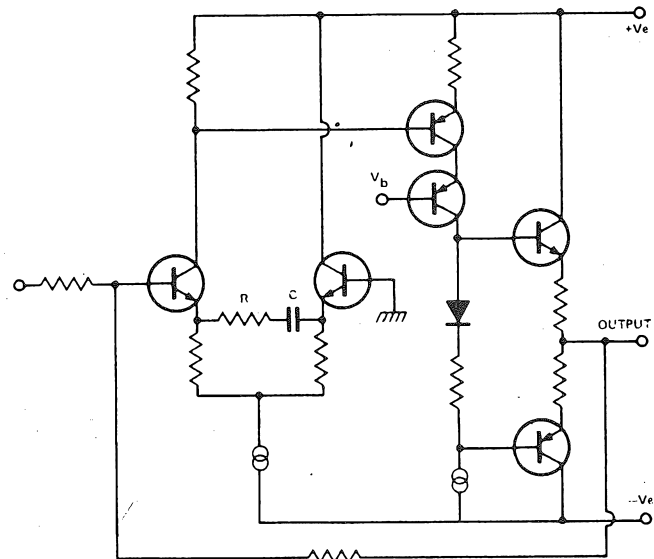


Fig 2. The other method. Lead compensation illustrated. Components R and C provide the time constant.

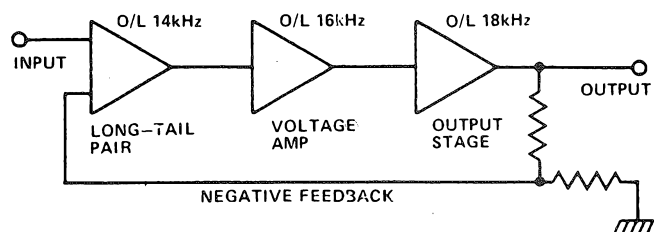


Fig 3. Third method of avoiding TID. Making each stage in the design have a wider B/W than the preceding one.

Benefits Conferred

An amplifier designed with a wide open-loop bandwidth, for low TID often has other, more tangible, benefits. The high frequency THD is usually no higher than at the mid-point; in stark contrast to more traditional designs. This is because gain is still available at high frequencies for negative feedback.

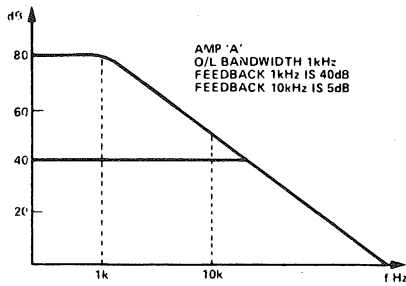


Fig 4. This amplifier design has a limited open loop bandwidth and the THD will rise with frequency.

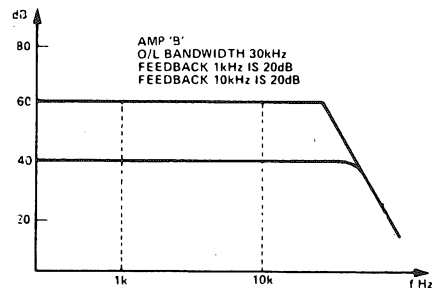


Fig 5. Contrast this with figure four above. The bandwidth here is much wider, resulting in a more linear THD response.

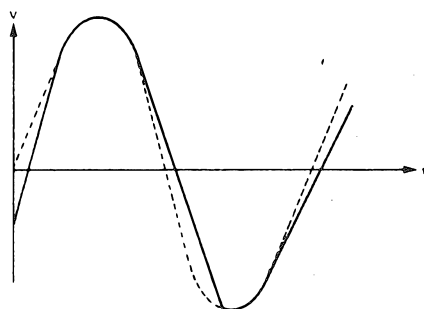
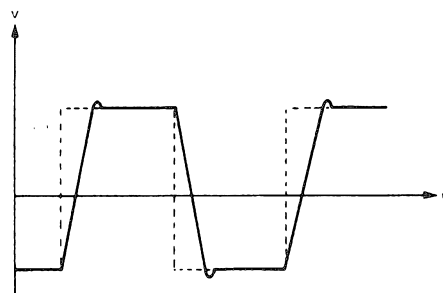


Fig 6. The effects of slow-rate limiting on a signal passing through an amplifier prone to this fault. Top: a squarewave, note the slight overshoot. Below: a sinewave. In both cases the dotted line represents the input.

Slew Who?

Such amplifiers also usually have much higher slew-rate. Slewing-rate defines the speed with which the amplifier can deliver output voltage to the load. For example, if an amplifier has a maximum output of 100 volts p/p and a rise-time of 10 uS, then the amplifier, if it were perfect, should have an output of about 80 volts

after 10 u secs in response to a suitable square wave input. In other words the output voltage would have risen at the rate 8V/uS.

However, amplifiers do not generally respond to large changes as fast as their small signal characteristics predict, for circuit and transistor capacitances can be charged only as fast as their driving circuits allow. In its simplest form the slew-rate of an amplifier defines how fast the output voltage can change for large signal conditions, and it is normally quoted in Volts per micro second. The maximum slew-rate of an amplifier is usually limited by the slowest stage in its circuit.

That stage will have an operating current I (as set in the design) and a capacitance C (usually a frequency compensation capacitor)

$$\text{Slew-Rate} = \frac{I}{C}$$

Thus if a transistor stage has a standing current of 100 u A and is compensated by a 33 pF capacitor then its Slew-Rate will be

$$\frac{100}{33}$$

i.e 3 V/u S.

Depending upon the design some circuits have a different Slew-Rate depending upon whether their output is negative-going or positive-going. Slew limiting also defines the full-power bandwidth; a figure more commonly quoted by manufacturers.

$$f_p = \frac{SR (10^6)}{2 \pi E_{op}}$$

$$E_{op} = \text{peak output swing in volts}$$

$$f_p = \text{Full power bandwidth in Hertz.}$$

Thus in a 100 Watt (into 8 Ohms) amplifier having full-power bandwidth of 20 kHz the required minimum slew-Rate would be about 5V/uS. This is, however, the absolute minimum figure and experience suggests that such an amplifier would have a hard, gritty high-frequency sound. Such an amplifier should have a Slew-Rate of greater than 20V/uS to be certain of avoiding the increase in distortion caused by the gradual onset of slew-limiting.

Unfortunately the higher the power output of the amplifier the greater the required slew-rate as more volts are swung at the output in the same period of time and so as our 100 W amp needs a 20V/uS an otherwise identical 50 W amp needs 14V/uS and a 20 W amp needs only 9V/uS.

Clip Around The Ear

But these forms of distortion tend to give subtle audible effects compared to the most common amplifier problem — that of clipping. Clipping occurs when an amplifier is overloaded by high level signal peaks. Such peaks occur frequently in much music material and so the manner in which the amplifier clips determines its audibility. A soft, clipping effect where the distortion rises gradually (typical of valve amplifier circuits) is audibly preferable to the hard clipping typical of transistor circuits.

Worse still, some amplifiers tend to suffer saturation

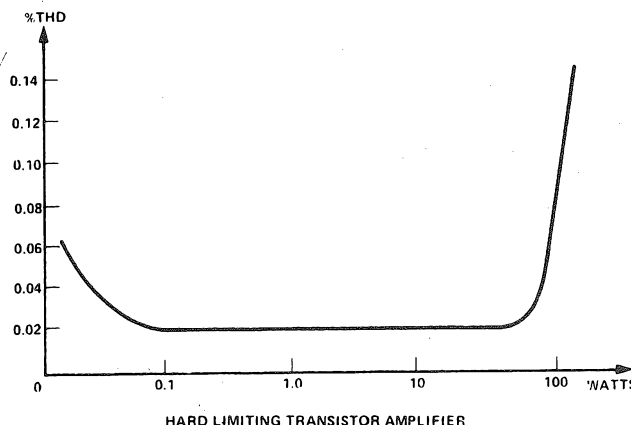
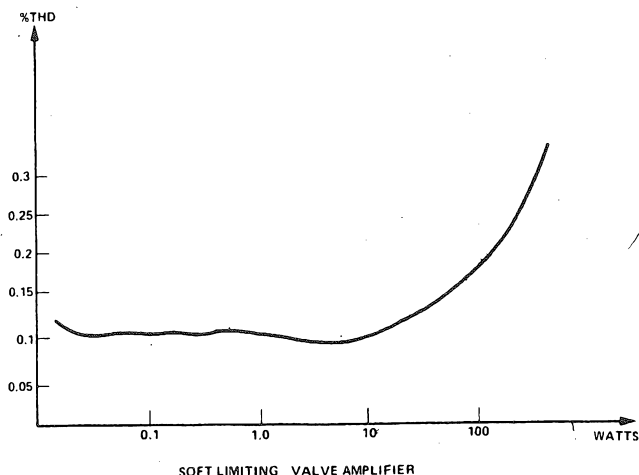


Fig 7. A comparison of the limiting characteristics — in general — of both transistor and valve amplifier types. There is a body of opinion which holds these curves to be the whole truth as to why valve amplifiers are preferred by many musicians.

effects on clipping, and take a time to recover; thus artificially extending the length of time the signal is clipped. The use of overall negative feedback to reduce distortion unfortunately makes things worse. Overall feedback effectively linearises the clipping making it hard; the distortion changes from 0.01% (say) to 10%, and quite suddenly too.

Designing A Designer

We have covered just a few of the requirements a designer must consider when working upon the design of power-amplifier. There are many more to be considered to even rough out a design specification before the circuit hardware is considered. The following sequence is mandatory:

1. What parameters are important to prevent any audible degradation of the signal?
2. Detail a performance specification that meets the requirements of (1).
3. Decide upon the circuit technology necessary; Bipolar; MOSFET; Valve; Class A; Class B; Switching; fast; slow; etc; etc.
4. Perform a development programme to produce a prototype.

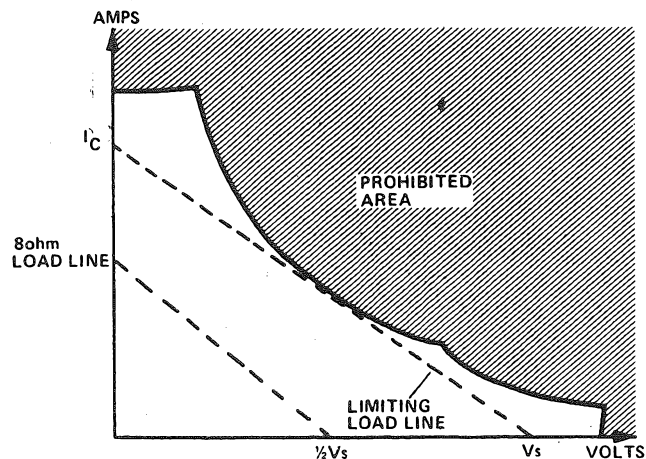


Fig. 8. Illustrating the load line conditions for output stages.

At this point the designer has to admit that it's a real world and that his performance specification cannot be achieved in a way that is acceptable to the accountants, salesman, customer, customer's wife or whoever else is around. Trade-offs are necessary and much to the "art" in amplifier design is in the deciding which defects and degradations are more acceptable than others.

As an illustration of the changes in design approach over the years we will briefly illustrate three designs for which the author has been responsible:

1. Cambridge Audio P60 (P80) (designed 1974)
2. Lecson AP3 Mk II (designed 1976)
3. Mission Electronics Voltage Amplifier (designed 1977)

Improvements

The P60 is capable of good mid-band performance (THD 0.01% at 1 kHz 30 W) but its high frequency distortion is poor because of the limited open-loop bandwidth. Generally this amplifier performs well at low and moderate levels but at high levels its sound quality becomes hard and aggressive. Some improvements to this circuit can be quite simply made as follows:

1. A resistor is fitted between Q10 collector and the negative rail to give better balance between Q8 and Q10:
2. A cascode transistor is fitted to Q13 collector to reduce "Early effect" distortion due to the collector-base capacitance of Q13.
3. An emitter resistor is fitted to Q13 to provide local negative feedback.

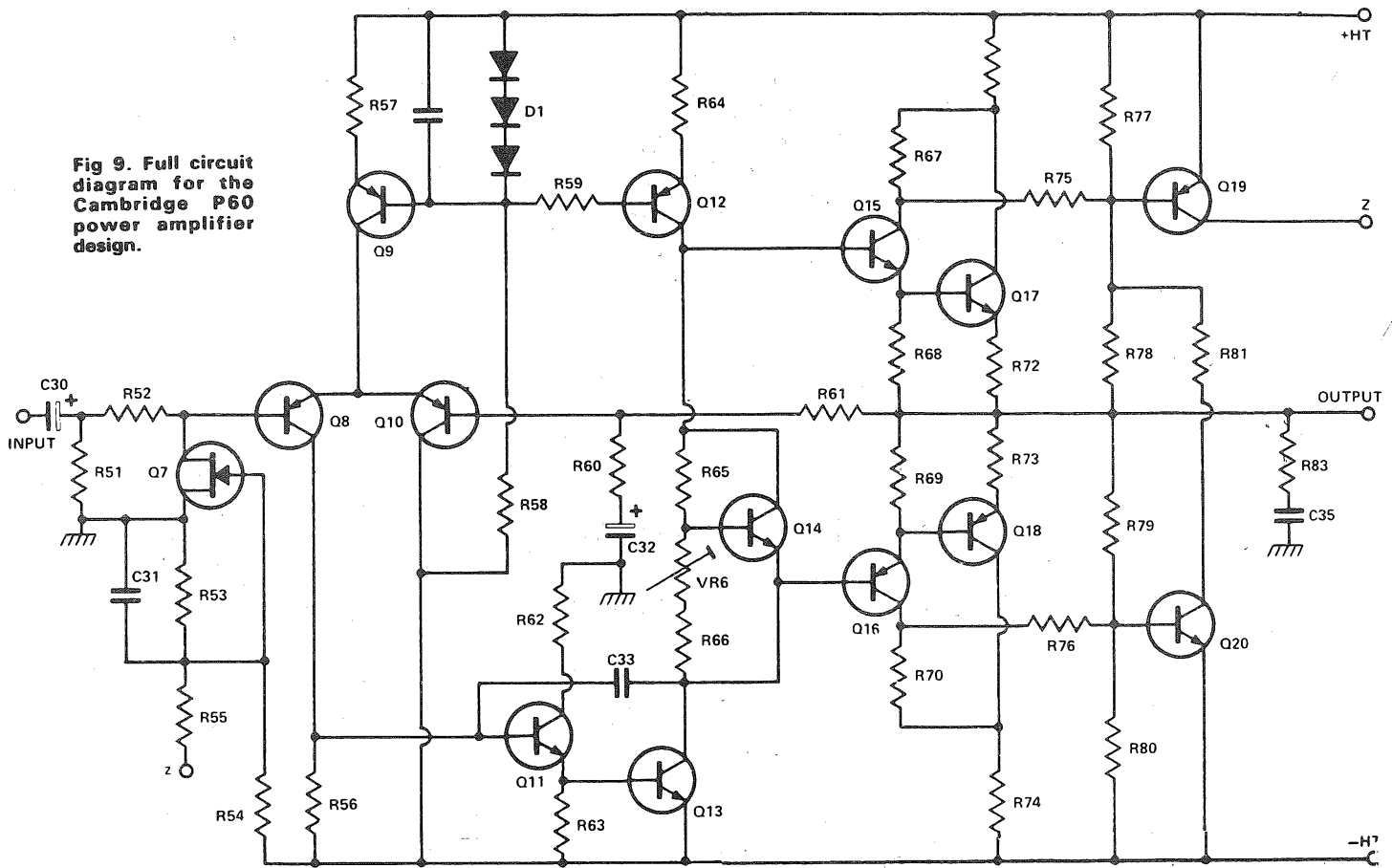
A Lesson Learned

The Lecson AP3 Mk II is an amplifier that incorporates much of the thinking in this article and is representative of the latest types of high performance amplifiers. It is a directly-coupled Class B design using a fully complementary output stage of series connected transistors and gives a power output of around 150 Watts per channel.

Mission Accomplished

The New Mission Voltage Amplifier represents an attempt to produce an amplifier that performs well irrespective of load. The circuits cannot be described at this stage as they are the subject of patent applications.

Fig 9. Full circuit diagram for the Cambridge P60 power amplifier design.



HOW IT WORKS—Cambridge P60

The P60 power amplifier is of a conventional design but with care being taken to optimise each stage. Q8 and Q10 form a long-tailed pair with Q9 as their emitter current source. Q8 and Q10 must be very closely matched for minimum DC offset and for maximum common-mode rejection to avoid H. T. ripple appearing at the output. The next stage is the Q13 voltage amplifier which is loaded by a current source (Q12) instead of the more common "bootstrapped" resistors. Note that Q13 is buffered

from the long-tail pair by an emitter follower (Q11) to prevent any loading of that stage worsening the distortion characteristics.

Capacitor C33 gives lag compensation which defines the dominant pole of the amplifiers. The open-loop bandwidth is quite high (for this type of circuit) at 12 kHz but none the less this amplifier is prone to TID effects. The protection circuit is very unusual in that the output is limited by an FET (Q7), Q19 and Q20 each form conven-

tional V-I summing circuits which monitor the loading on the output stage.

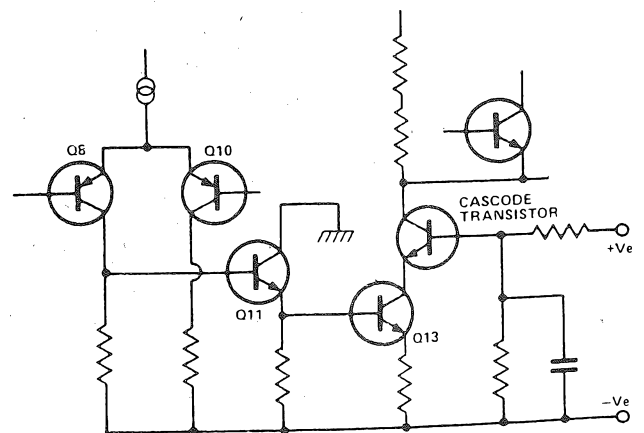
If either Q19 or Q20 turns-on, the gate of the FET Q7 (normally biased-off by R54 to the negative HT) is biased positive and it starts to turn-on. It then acts as a potential divider with R52 and thus attenuates the audio signal. This protection only turns on at the equivalent of 50 W into 2 Ohms load and when it turns on it only adds moderate distortion (0.2% typically) as distinct from clipping.

However, a brief description will illustrate the philosophy behind the design.

The casing contains two completely separate Mono Amplifiers each with its own power supply. A separate module carries the DC-voltage offset protection circuits; the delayed switched-on circuits; and the thermal protection circuits. Particular attention has been paid in the design to achieving:

1. Low distortion with a very low order of overall feedback
2. Wide open-loop bandwidth with an excellent slewing rate
3. Minimum time and phase distortion
4. A high transient power capability with virtual freedom from clipping effects.

Fig 10 (Right). Showing how some of the improvements mentioned can be added to the P60 basic design.



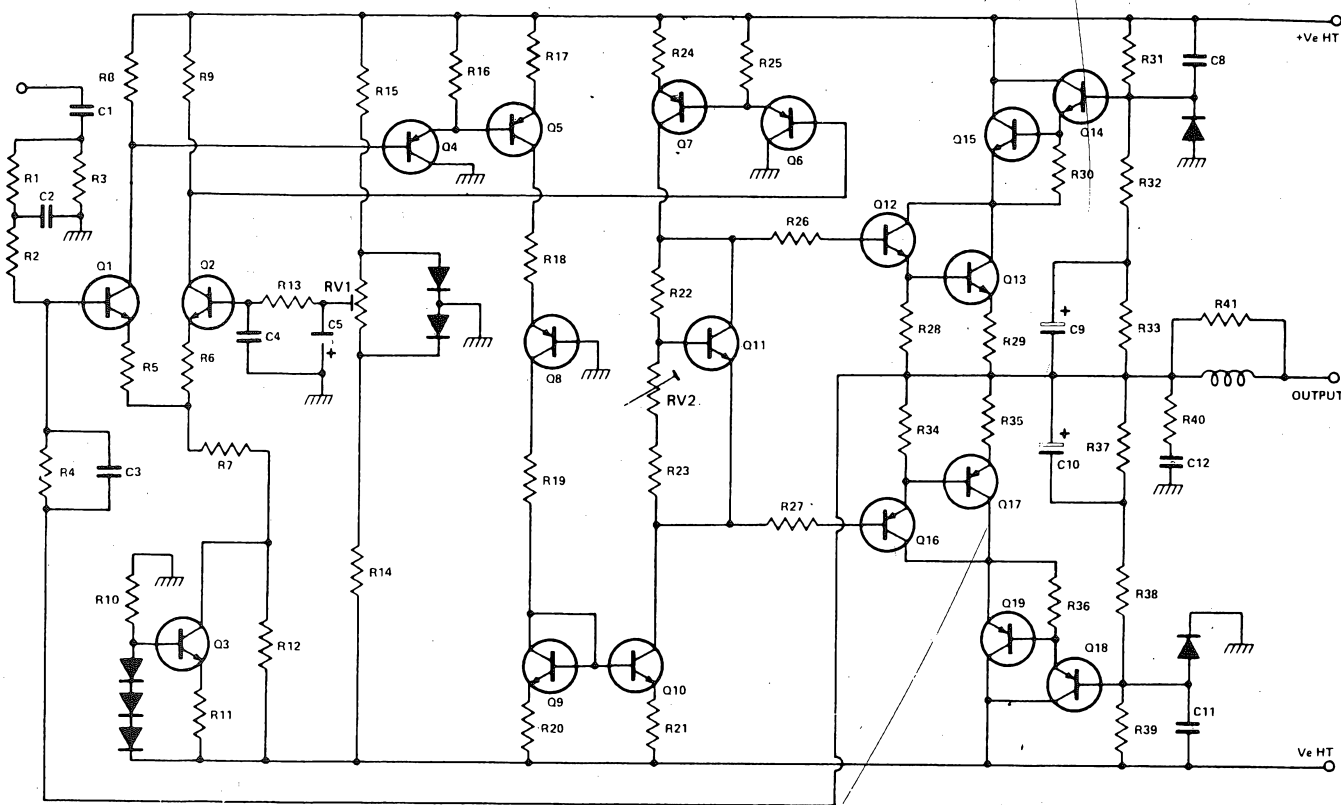


Fig. 11. Full circuit diagram for the Lecson AP3 power amplifier design, producing around 150W.

HOW IT WORKS—Lecson AP3

Transistors Q1 and Q2 form a long-tailed pair differential amplifier with Q3 as the emitter current source. Local feedback is applied in the form of emitter resistors R5 and R6. The base of Q2, instead of being grounded, is connected to a potential divider RV1 which permits the DC offset at the output to be set to zero. The input signal to Q1 is passed through a low-pass filter (R1, C2) which sets the bandwidth to 22 kHz (i.e. below the open loop bandwidth for no TID effects). The bi-phase outputs of the long-tail pair feed a second differential amplifier Q5 and Q7. Transistor Q5 has a constant current load (Q8) whilst Q7 is terminated by a current mirror (Q9 and Q10). Transistor Q10 will always deliver the same current as transistor Q9 hence the term "Current Mirror" and the excellent symmetry and balance this stage achieves. Functionally, however, Q10 can be considered as an active load whilst Q7 is a voltage amplifier from whose collector the drive to the output stage is taken. Note that Q5 and Q7 both have local emitter feedback (R17, R24) and that both are buffered from the long-tail pair (Q4 and Q6 emitter followers).

Transistors Q12, Q13, Q16 and Q17 each form conventional Darlington emitter follower stages. Each stage is series connected to a further power transistor (Q14, Q15 and Q18, Q19 respectively) which is permanently biased ON. Their emitter potentials are determined by the ratio of the base potential dividers. This ratio was chosen such that Q13 and Q15 each has half the supply rail across them.

The whole amplifier is in the inverting mode with overall shunt feedback through R4 and C3.

This amplifier is quite fast having an open-loop bandwidth of about 27 kHz. The circuit is stable without the usual compensation capacitors within the loop. THD is low being typically (at 100 W into 8 Ohms) 0.004% at 1 kHz and 0.02% at 10 kHz. The HF distortion can be further improved by selection of transistor Q7 for a device with a low collector-base capacitance.

No conventional protection circuits are used as extremely high power transistors are fitted and these can survive a short-circuit condition in the time taken for the power supply to shut down.

The output stages have a very high current capability but have no protection circuits, the output transistors being designed to sink the full energy of the power-supply into the load. A patented form of voltage feed to this stage gives the amplifier a short term power delivery capability of about 600 Watts (compared to the rated 150 Watts 8 Ohms). This, of course, represents a 6dB increase in power availability over the rated figure. The voltage amplifying stages are designed to clip softly and this combined with the low-overall feedback gives the amplifier overload characteristics similar to those of an equivalent valve amplifier.

Conclusion

This article had discussed just some aspects of modern audio amplifier design. Much attention is focused upon whether an amplifier is designed around bipolar transistors, FETs, valves, or switching transistors. Designers are now finding that the major stumbling block is not designing a circuit using any of these technologies but in deciding upon what is the performance specification required that will give faithful reproduction of the sound source. Until this problem is solved there will continue to be an element of uncertainty in amplifier design.

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INTO ELECTRONICS PLUS

INTO ELECTRONICS PLUS has one hundred pages containing the whole of Ian Sinclair's series Into Electronics, PLUS a selection of features and articles that cover subjects ranging from Computers to Capacitors.

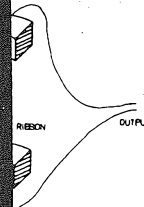
to used in some types of form, a ribbon is suspended powerful magnets. Leads are bottom of the ribbon. ribbon, it will vibrate and the magnet and a current will be connected wires.

MISTER SPEAKER

It would be logical to assume that if microphones produce an electrical signal when excited by sound, they should produce sound when an electrical signal is applied.

The theory behind this does in fact support this assumption to all the microphones mentioned except the carbon microphone.

In practice however, the requirements of an electric to sound transducer are such that mechanical design of loudspeaker and earphones is different.



the ribbon is very small. Any is connected would have a produce the signal quite consid- transformer is always mounted the transformer increases the of the ribbon's output so that less attenuating effect. For of microphone is called a

ry fragile and has to be well heavy breaths) making it If a coil of wire were used different mechanical assembly) n higher. This is done in the hone. A diaphragm is attached in the magnetic field. The If this type of microphone is re is no longer a thin vulner-

he diaphragm and coil is much e ribbon. At high frequencies rder for the sound to vibrate it

device more rugged other The poorer high frequency come by making the coil and ter. This would unfortunately sponge

some microphone manufac- phones into one body. One the other a much smaller one es however, are much more

used by Bell was one of the nd not the carbon type by the as very low, and as amplifiers d not reproduce at the receiving

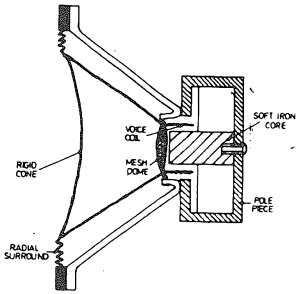


Fig. 7.1 Cutaway of a moving coil loudspeaker

The moving coil loudspeaker consists of a paper cone attached to a fine coil of wire, which is suspended in a strong magnetic field. The edge of the cone is attached to the metal frame of the speaker so that it is free to move. When an electric signal is applied to the coil (called the voice coil) a magnetic field is set up around the coil. This interacts with the magnetic field caused by the large magnet and the coil is moved. The coil and thus the cone move in sympathy with the signal in the voice coil.

As with microphones (only more so), large loudspeakers with big cones are only capable of effectively reproducing sound in the lower end of the sound spectrum. For this reason, all high fidelity loudspeakers contain two or more 'drive' units. Each unit will be of a different size and designed to handle a particular range of audio frequencies.

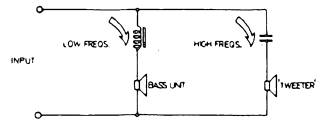


Fig. 7.2 A simple crossover

medium waveband isn't enough room. Various intermediate frequencies above 5 kHz can't try to ensure spaced frequency schemes could have which choose to them, and of cou of profit in ruinin has now made t serious broadcast

circuits, is it? We get around the pnc the circuits have built-in circuit excessive voltage damage can occur only different voltage from a connected into its circuit nected through resistors voltage, and the voltage b than the battery voltage. v problem, then, is just to g one piece CMOS ICs c embedded in plastic loan conducting material that r shorted together, keeping Keep the ICs in this and lightning

DO NOT TOUCH

Now for the awkward bit circuit. Well, we build a checking to make sure it connected to a part of the connected, or is earthed, there two ways of dealing IC holders into the circuit directly. We then hold th fingers away from the pl and plug in the ICs, marki right way round. Unless we are able to do this without touching any of the pins the other method is direct soldering. The negative line on the board is connected to earth through a flexible wire fitted with croc clips. The ICs are fitted, one at a time into their places, and the pins are soldered in one at a time using an earthed soldering iron. The earth pin of the IC is soldered in first, then the + supply pin, then all the others. This is no more trouble once you have had a bit of experience, but the less experienced constructor is

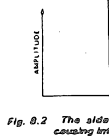


Fig. 6.2 The slide causing link

to keep the load is not continuous steady; from this lot is there a sine wave, whose f oscillator

DAH, DAH D

So far, so good. We will radiate from a w How do we use it?

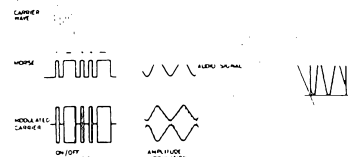


Fig. 7.2 Amplitude modulation of a carrier wave.

knew to use a radio wave was the way we used the current in a telegraph wire — we started and stopped it in the pattern called Morse code. Now the Morse code is

did not yet exist the sound reproduced at the receiving end very faint

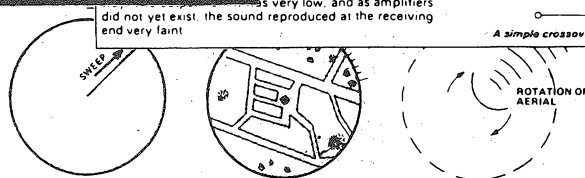


Fig. 6.7 PPI Trace (a) direction and rotation of sweep on screen (b) typical trace, showing transmitter at centre and the surroundings, with stationary objects and moving targets. A long persistence tube is used, so that the targets leave trails to indicate their movement.

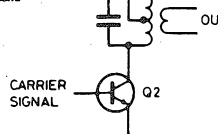


Fig. 7.3 AM by means of an emitter follower.

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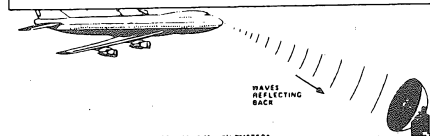


Fig. 6.17 The seven-segment display.

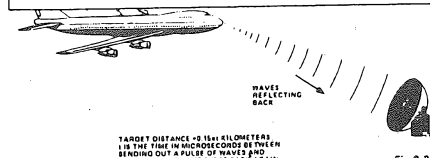


Fig. 6.3 Principle of radar.

ETI

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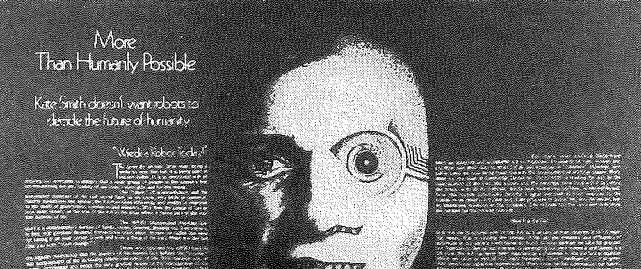
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PRACTICAL GUIDE TO TRIACS

Triacs are simple, versatile devices used throughout electrical and electronic engineering. Yet most electricians regard them with almost superstitious awe. This three-part article describes their uses in a totally practical way.

Electronics, according to a friend of ours, is anything electrical that he doesn't use or understand.

Triacs, by this definition, must soon be considered non-electronic, for they are finding ever-increasing use throughout electrical engineering — in both consumer and industrial applications.

In effect, a triac is similar to a latching relay. It closes — practically instantaneously — after being triggered, and remains closed until the supply voltage is reduced to zero (or changes polarity). When this occurs (twice every complete cycle if used on ac) the triac opens, but will close again almost instantaneously if retriggered.

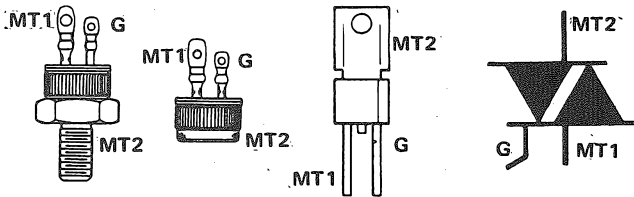


Fig. 1. Construction and electrical connections of typical triacs.

Static switching

Triacs can usefully replace mechanical switches in ac circuits. They allow the control of relatively high power by very low triggering current — and as triacs latch each half-cycle there is no contact bounce; nor, as they open only at current zero, is any arcing or transient voltage developed due to stored inductive energy in the load or power lines. They eliminate completely the contact sticking and wear associated with electro-mechanical relays, contact-breakers, etc.

Triggering arrangements are very flexible — most triacs can be switched into conduction at any point on either half-cycle of the ac waveform by applying a low voltage of either polarity between the gate electrode and main terminal 1. (A few types of triac can only be switched by a negative-going gate voltage or pulse.)

The triggering voltage can be obtained from a battery (Fig. 2) or simply from the ac mains (Fig. 3). In either case, full wave current will flow when SW1 is closed and current will cease to flow at the end of the half-cycle (whether positive or negative) in which SW1 is opened (Fig. 4).

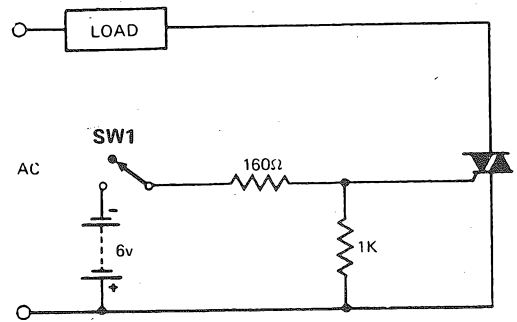


Fig. 2. Triac triggered by external dc voltage.

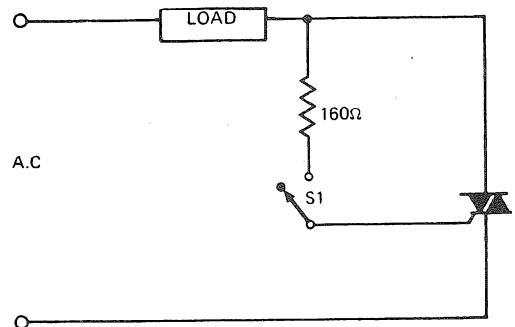


Fig. 3. Triac is triggered by input ac, in this application SW1 will be live.

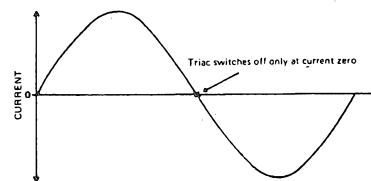


Fig. 4. A triac can be triggered into conduction at any point along the sine-wave, but will only switch off at current zero.

Eliminates burnt contacts

The simplest possible method of triggering is shown in Fig. 3. This circuit is often used to eliminate burnt contact breaker points in thermostats and similar devices which have to make and break large currents at frequent intervals. The existing make and break arrangement is retained but is used only to switch the triac, which in turn switches the main load current. The current flowing through the contacts is reduced to a few milliamps.

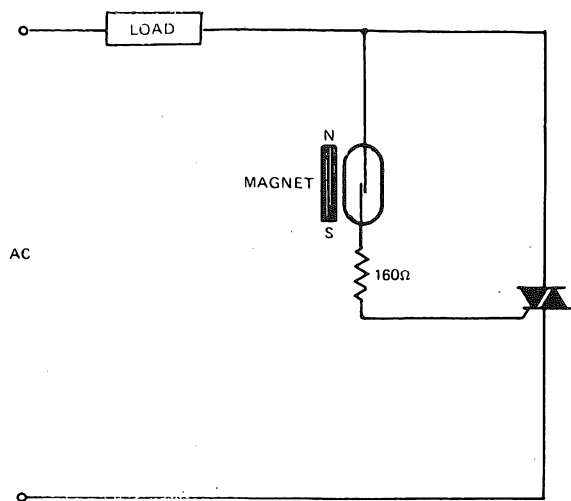


Fig. 5. A magnetically operated reed switch provides electrically isolated triggering from mechanical movement. The switch will close when the magnet is brought within half-an-inch or so.

Movement switches triac

A magnetic method of triggering may be used when a mechanical movement actuates an electrical circuit. To do this, a magnetically operated reed switch is used as SW1 (Fig. 5). The switch contacts are closed when a magnet is brought near the switch. The actual distance will depend upon the sensitivity of the reed switch and the strength of the magnet ($\frac{1}{2}$ " to 1" is typical). Reed switches were more fully explained in Electronic Circuit Design No. 2.

Reed switches may also be used to provide electrical isolation between the triac and the triggering circuit. The reed is inserted in a coil which is then switched by a suitable low voltage dc supply (Fig. 6). As the life of reed switches exceeds several million operations, this is an extremely reliable method of switching.

Other simple methods used to isolate the triggering circuit from the triac are shown in Figs. 7 and 8. The opto-isolator shown in Fig. 8 provides extremely high electrical isolation. Opto-isolators, in which a light source and photo-cell are integrally mounted, are commercially available for as little as a pound.

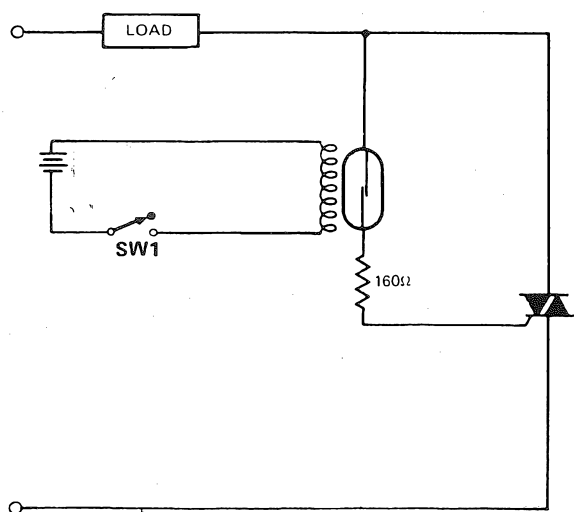


Fig. 6. Reed relay can be electrically operated to provide electrical isolation.

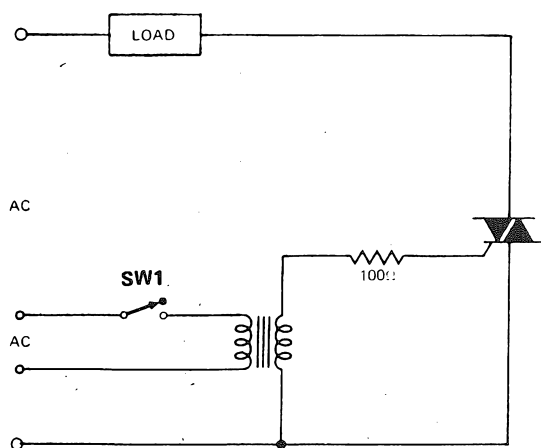


Fig. 7. Isolated ac provides triggering.

An unusual off/half-power/full-power circuit is shown in Fig. 9. When half-power is required, the diode is switched in series with the triggering lead. This causes the triac to conduct only on alternate half-cycles. The circuit is suitable for heating, or other resistive loads that have thermal inertia. It is not suitable for lighting control, as the halved frequency will cause an irritating flicker, nor should it be used for inductive loads such as motors or transformers.

A latching circuit is shown in Fig. 10. Momentarily depressing SW1 will cause the triac to conduct and to remain conducting after SW1 is released. The circuit is reset by momentarily depressing SW2; C1 should be a $0.5 \mu\text{f}$ to $2.0 \mu\text{f}$, 630 volt working non-polarised capacitor.

Time delay circuits

Triac time delay circuits are shown in Figs. 11, 12 and 13. All three circuits will provide time delays up to 100 seconds or so and may be used for applications such as photographic timers, industrial machinery, etc.

In the circuit shown in Fig. 11 the triac will conduct as soon as SW1 is closed. After a time — determined by the setting of the one Megohm potentiometer — the uni-junction will fire, causing SCR1 to conduct thus removing the triggering voltage from the triac.

Other types of time delay circuits are shown in Figs. 12 and 13. In both of these circuits the triac will conduct a predetermined time after SW1 is closed. The preset time intervals are set by the 1 Megohm potentiometer.

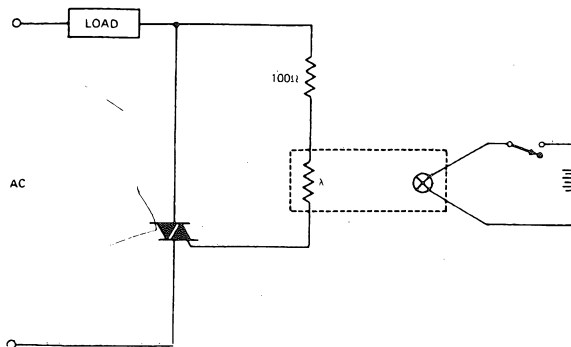


Fig. 8. Photocell coupling provides electrical isolation.

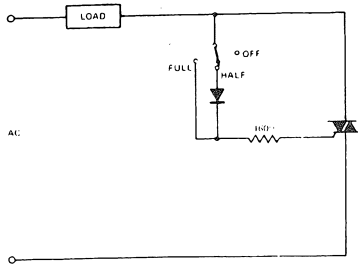


Fig. 9. Circuit can be switched to provide half or full power into load.

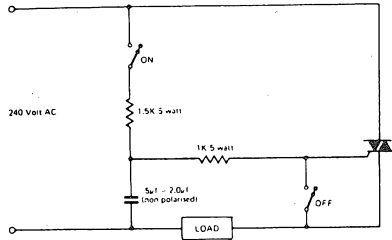


Fig. 10. This circuit provides latching action.

Liquid level is used to switch the triac circuit in Fig. 14. The triac will conduct when the resistance between the sensing probes falls below 100K. The pulse transformer in this circuit must be well insulated to ensure that mains voltage cannot appear across the probes.

Inductive loads

All the circuits described so far will operate reliably with resistive loads. Inductive loads present a minor problem. For reliable operation it may be necessary to limit the rate of voltage rise immediately following turn-off. This is done by connecting a capacitor and series resistor across the triac (Fig. 15).

A triac's power rating is a function of its operating temperature. Thus a unit of 6 amps nominal rating used without a heat-sink must be derated to 1½ amps. A rough guide can be obtained from the triac's external case temperature. This should not exceed 75-80°C.

In the majority of circuits the outer case of a triac will be 'live'. Because of this it is necessary either to insulate

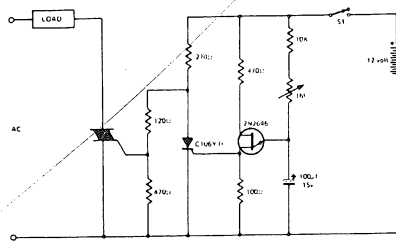


Fig. 11. Time delay circuit — power is disconnected from load after preset time.

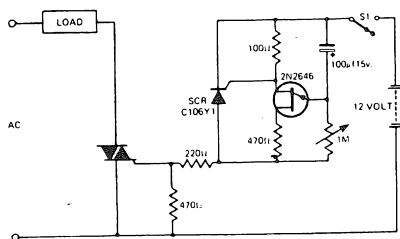


Fig. 12. Time delay circuit — power is connected to load at end of preset delay.

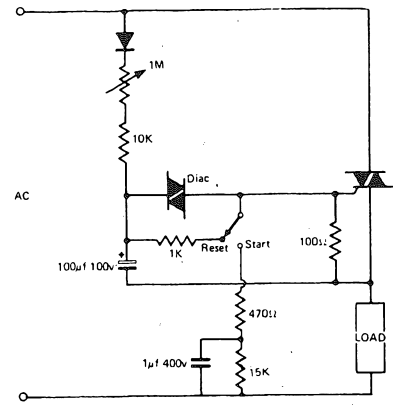


Fig. 13. Time delay circuit — power is connected to load after a preset time.

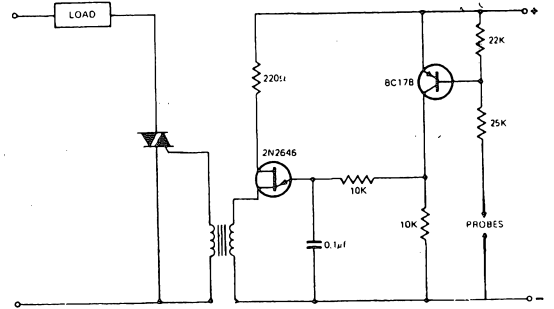


Fig. 14. Liquid level switch — triac is switched on when liquid covers probes.

the heat-sink from surrounding metalwork, or to insulate the triac from the heat-sink.

Two flat mica washers and a Teflon collar will effectively insulate the triac from the heat-sink, but there will be an appreciable loss in cooling efficiency. Where possible, it is better to have metal-to-metal contact and then to use an insulated mounting for the heat-sink itself.

Providing triacs have adequate heat-sinks, it is permissible to run them permanently at their maximum designed rating. They can also withstand considerable overloads for short periods. A typical unit can withstand ten times its nominal rating for one half-cycle of a 50 Hz waveform.

Do not overload

Despite this tolerance of short-term overloads, triacs must not exceed their designed ratings for more than a fraction of a second. A short circuit will destroy them instantaneously. Fusing is usually ineffective, as a triac will almost invariably fail first and protect the fuse.

An almost certain symptom of overloading is loss of gate control; the triac will appear to have a short circuit from one main terminal to the other, although overloaded triacs will occasionally still operate on one half-cycle only.

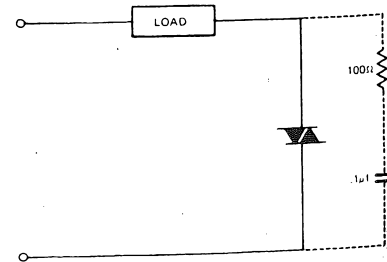


Fig. 15. Capacitor and resistor should be connected across the triac when switching inductive loads.

PRACTICAL GUIDE TO TRIACS

PART 2

The second article of this three-part series describes how triacs are used to control the amount of power applied to electrical loads.

Triacs not only switch power on and off and thus replace mechanical contacts — they can also control the *amount* of power applied to a load.

These are two quite different ways of using triacs for this purpose — 'phase control' and 'zero voltage switching'.

Phase control is the method used to control light dimmers, electric motors, and for voltage and current regulators.

Heating loads can also be controlled in this way — but, for reasons outlined in the third article, zero voltage switching is a better technique.

How phase control works

A triac can be triggered into conduction during any part of a half-cycle of applied voltage, and will *remain conducting* until the end of that half-cycle.

Thus, if a triggering signal is applied at the beginning of each half-cycle, the triac will in effect conduct as if it were a switch that had been turned on. But if the triggering signal is applied halfway along the waveform, the triac will conduct only for the remaining period of that half-cycle; and as the triac conducts for only half the time, it reduces the power applied to the load by roughly the same amount (Fig. 16).

Thus the power input to the load can be set at any level simply by controlling the point at which the triggering signal is applied. Oscilloscope pictures (Figs. 17a, b, c) show the 50Hz waveform triggered at three different points. Notice the corresponding change in brilliance of the lamp used as a load.

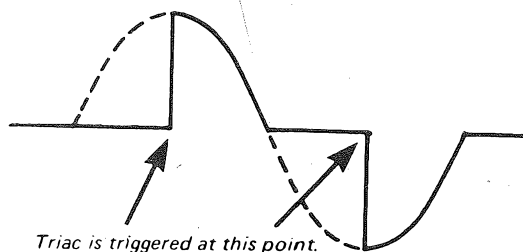


Fig. 16
Drawing shows Triac triggered at half-way point.

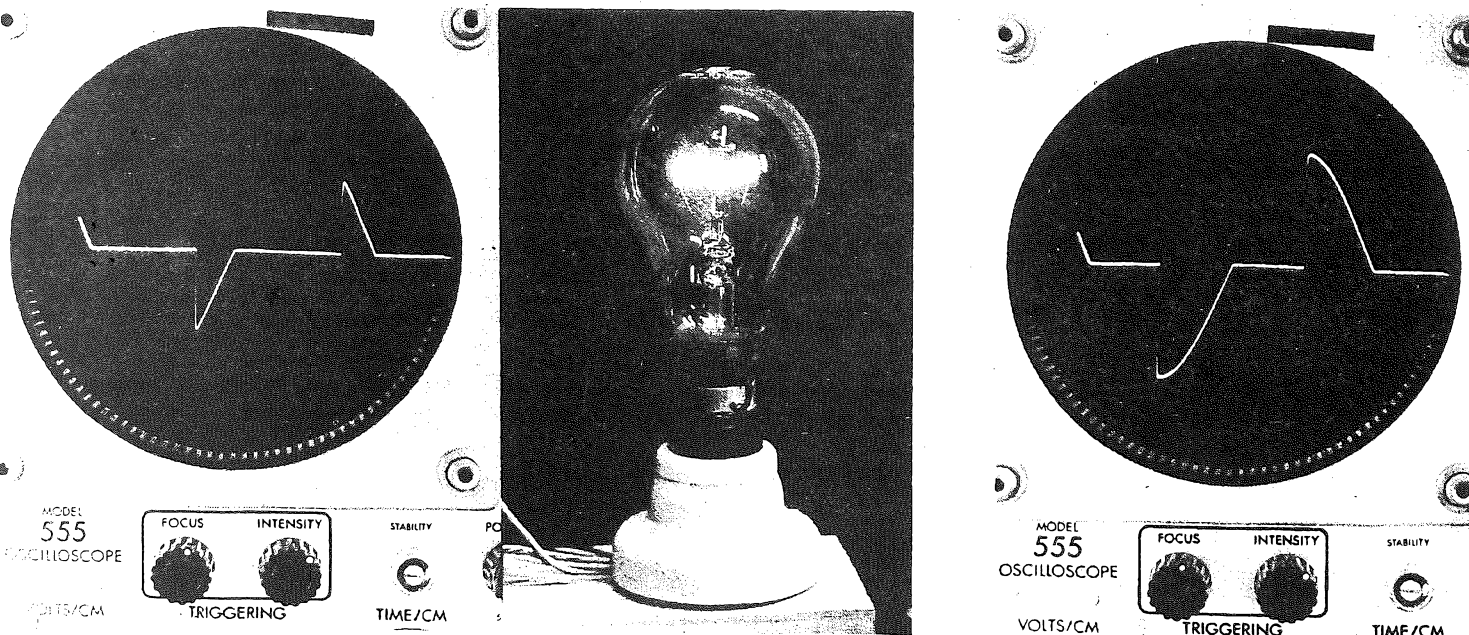
Light dimmers

The circuit of a commercially produced light dimmer is shown in Fig. 18.

During each half-cycle the 0.1 μf capacitor charges up (via the control potentiometer) until it reaches 30-32 volts. At this voltage the diac (trigger diode) conducts and the voltage across the capacitor gates the triac into conduction. (A neon lamp is sometimes used instead of a diac).

The time required for the 0.1 μf capacitor to charge to the diac triggering voltage is determined by the resistance of the control potentiometer. When this is reduced to zero, the capacitor will charge to the trigger voltage instantaneously and the triac will be gated into conduction for practically the complete half-cycle. At maximum potentiometer resistance the capacitor may not reach the diac triggering voltage until the end of the half-cycle, and the triac will conduct for only a very short time.

Fig. 17
Actual waveforms of a domestic light dimmer showing triggering at $\frac{1}{4}$, $\frac{1}{2}$ & $\frac{3}{4}$ points. Note corresponding changes in lamp brightness.



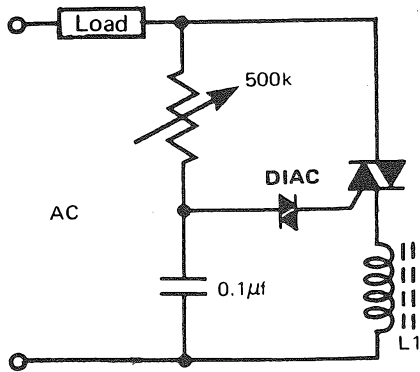


Fig. 18
Basic phase control circuit.

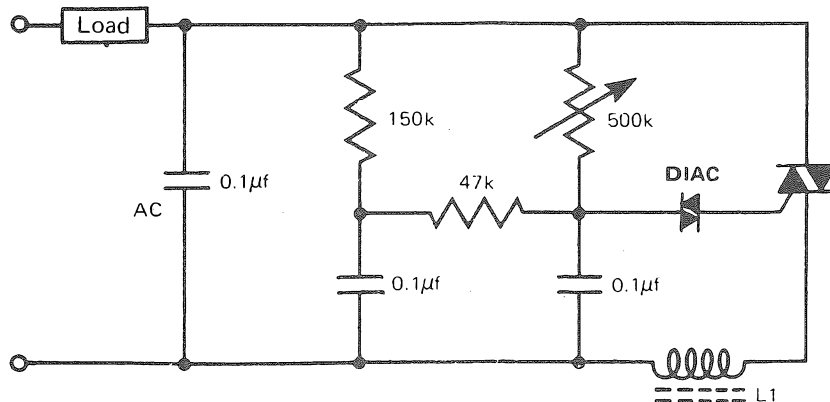


Fig. 19
Simple versatile phase control circuit is suitable for domestic light dimming and other resistive loads.

The circuit shown in Fig. 18 is very simple and cheap to make but has one major limitation namely, that power cannot be controlled smoothly from zero to maximum. As the potentiometer control is advanced, the load current will jump quite suddenly from zero, to some intermediate value from which it can only then be controlled smoothly in either direction.

If the power supply is momentarily interrupted while the lights are dimmed below this 'jump' (hysteresis) level, the lights will remain off when power is subsequently restored.

Reducing hysteresis

The hysteresis effect is considerably reduced in the circuit shown in Fig 19. This circuit is ideal for use as a domestic light dimmer. All components can be mounted behind a wall switch plate — and if the load does not exceed 200 watts, the triac will not require a heat sink.

Almost total freedom from hysteresis is required for theatrical light dimmers. This can be achieved by using the circuit shown in Fig. 20.

What size triac?

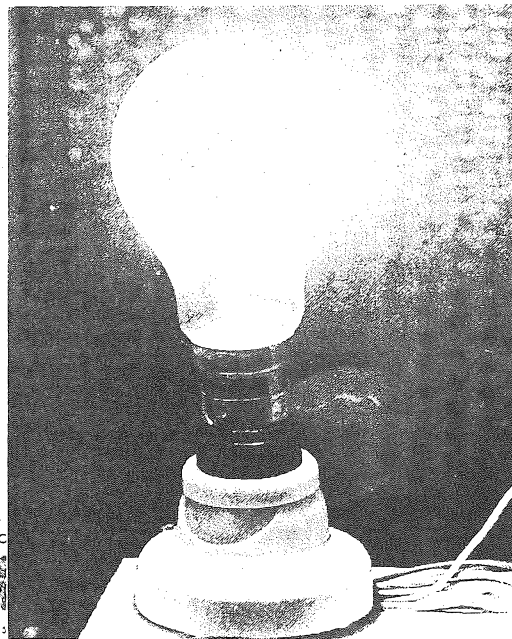
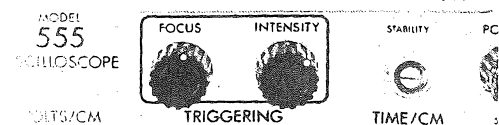
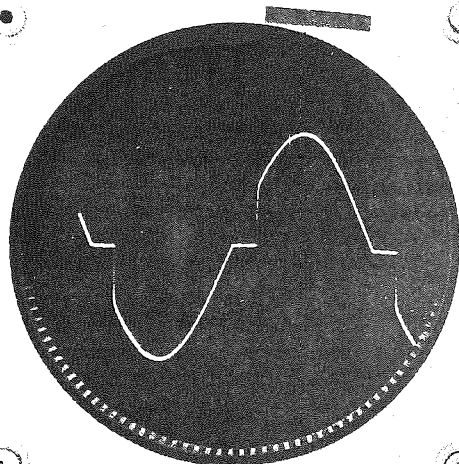
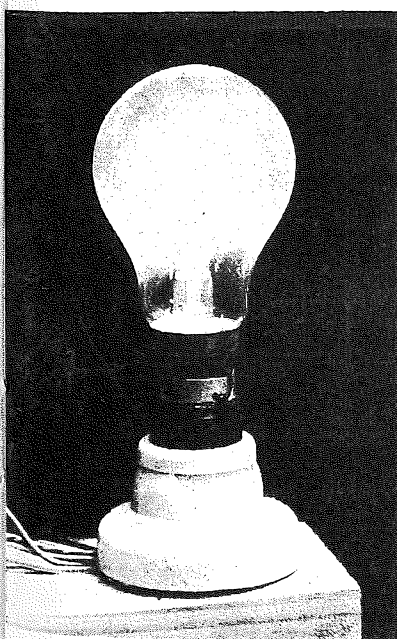
Incandescent lamps draw a very heavy current during the time taken for the filament to reach operating temperatures. This on-rush current may exceed the working current by 10 to 12 times. Fortunately domestic light bulbs reach working temperature within a cycle or two, and the short-term overload is absorbed by the triac.

This does not apply to theatrical lighting equipment, where the larger wattage bulbs take longer to reach working temperature. For this type of use the triac must be rated to carry at least five times the normal maximum load.

Voltage fluctuations

All the control circuits shown so far are voltage dependent — that is, the output voltage changes if the input voltage changes.

This voltage dependence can be overcome by using a circuit in which a zener diode plus a dc error signal compensates for changes in line voltage. (Fig 21).



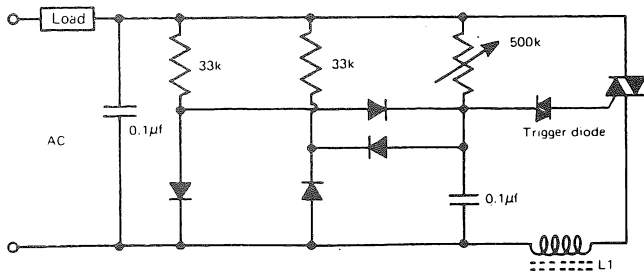


Fig. 20
Suitable for theatrical light dimming, this circuit has very low hysteresis.

This arrangement will hold the rms output voltage within 5% for a 50% change in line voltage. It is often used in photographic and other applications where a constant light level is required.

Fluorescent lamps

With all the phase control circuits shown so far incandescent filament lamps can be controlled with no further modifications to the house lighting system. Fluorescent lamps can be dimmed but they require special circuitry — damage will occur if the circuits shown here are used.

Quartz iodine and similar lamps using a regenerative halogen cycle should not be dimmed for any length of time. If the envelope temperature of the lamp drops below 250°C, the regenerating halogen cycle ceases. Tungsten from the filament will be deposited on the wall of the lamp, reducing filament life and decreasing the light transmission of the glass.

A modification which can be used with any of the circuits discussed so far is shown in Fig. 22. This arrangement turns lights on as darkness falls and turns them off again at daybreak. The photo cell must see ambient light and be shielded from the controlled light.

Motor speed control

Triac phase-control can be used to vary the speed of electric motors.

The universal type of series-wound motor may be controlled by circuits similar to those used for light dimming — except that, to ensure reliable commutation, a capacitor and series resistance must be connected in parallel across the triac (Fig. 23). With this arrangement the speed of the motor will vary with changes in load and supply voltage, but for non-critical applications (such as fan speed control), where the load is constant at any given speed, the circuit may well be adequate.

Motor speed which, once pre-set, is maintained despite load changes is a useful characteristic for power tools, laboratory stirrers, watchmakers' lathes, potters' wheels etc. To obtain this 'load sensing' characteristic, an SCR is commonly used in a half-wave configuration (Fig. 24). The circuit works quite well over a limited range of speed but is prone to low-speed jerkiness and the half-wave operating principle precludes controlled operation much beyond half-speed.

A load sensing phase-controlled circuit in which a triac provides full zero to maximum control is shown in Fig. 25.

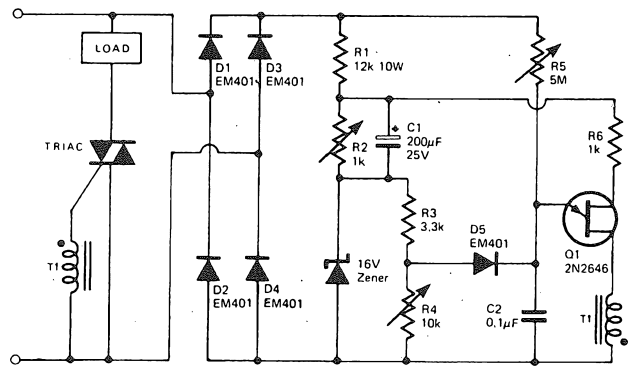


Fig. 21
Zener stabilised circuit maintains constant pre-set output despite changes in line voltage.

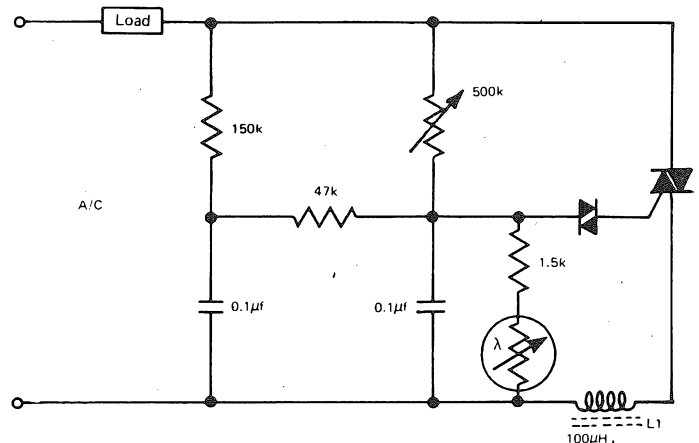


Fig. 22
Photo-cell switches Triac on at night, off during day.

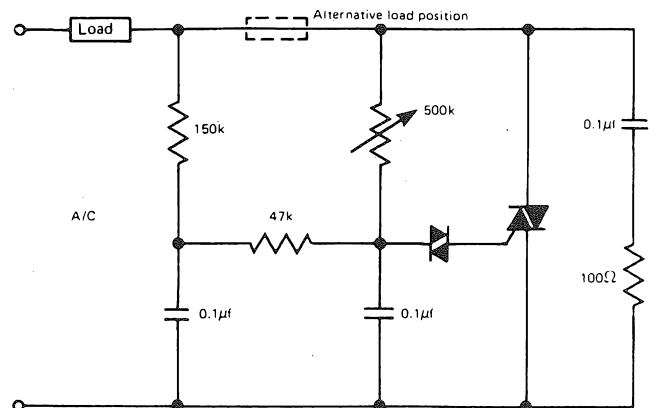


Fig. 23
Phase-control circuit for inductive loads, it may be necessary to connect the load in alternative position shown to obtain adequate control.

Induction motors

Induction motors can also be triac controlled — but some problems may be encountered, especially with split-phase or capacitor start motors.

In general, induction motors can be controlled between full and half-speed, providing they are not run at full load. A reasonably reliable guide is motor operating temperature. This must not exceed the manufacturer's rating at any speed.

Once again, the modified light dimmer circuit shown in Fig. 23 can be used, but the load should be connected in the alternative position shown in dotted lines.

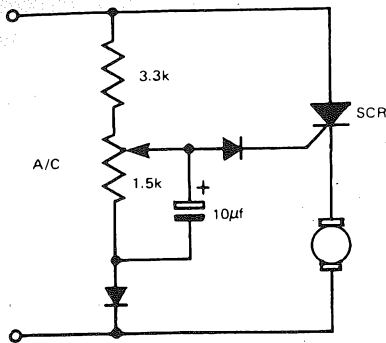


Fig. 24
Half-wave SCR circuit provides feedback speed control.

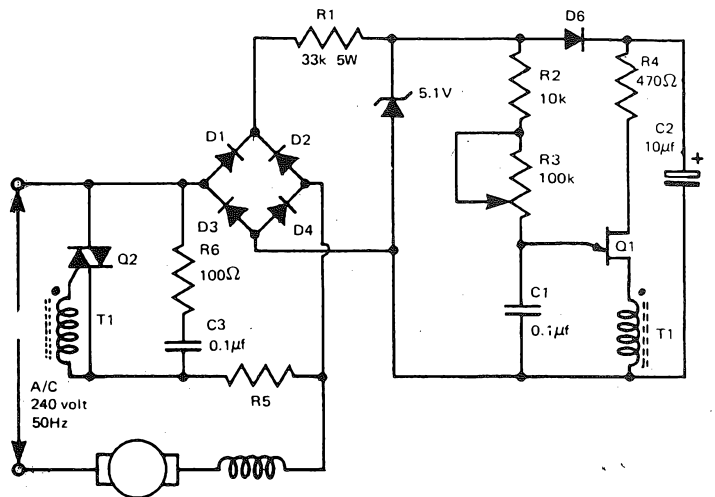


Fig. 25
This full-wave load sensing circuit will control ac/dc motors over wide speed range.

Varying transformer voltage

The circuit arrangement described above can also be used to vary the voltage in the primary winding of a transformer and thus obtain an adjustable secondary output.

This arrangement was used in a range of microscope lamp controllers designed by the writer some years ago. An adjustable zero-set was obtained by replacing the 4.7k resistor by a 100k potentiometer.

Heating loads

Any of the triac phase-control circuits shown so far can be used to control heating loads, although the temperature of the controlled load will vary with changes in line voltage and ambient temperature.

A circuit which compensates for these changes is shown in Fig. 26. Theoretically this circuit can maintain temperature to within 1% of the set point despite line voltage changes of $\pm 10\%$. Actual performance depends upon the design and construction of the system in which the controller is used.

This circuit provides proportional control — that is, full power is applied to the heating elements while the load is warming up, until, at some intermediate state, the power is reduced by an amount proportional to the difference between the actual load temperature and the required load temperature. The proportional band is adjustable by a 'gain' control.

The circuit is simple yet effective, but it has one major failing which restricts its use to all but light loads. This is radio interference.

Radio interference

All phase control systems generate large amounts of radio interference, primarily at low and medium frequencies. It seriously affects the long and medium wave reception of radio transmissions and will also be picked up by audio equipment, when it is heard as a raucous buzz. This rfi can also interfere with laboratory apparatus, especially pH meters, and has been known to cause erratic operation of digital computers and other

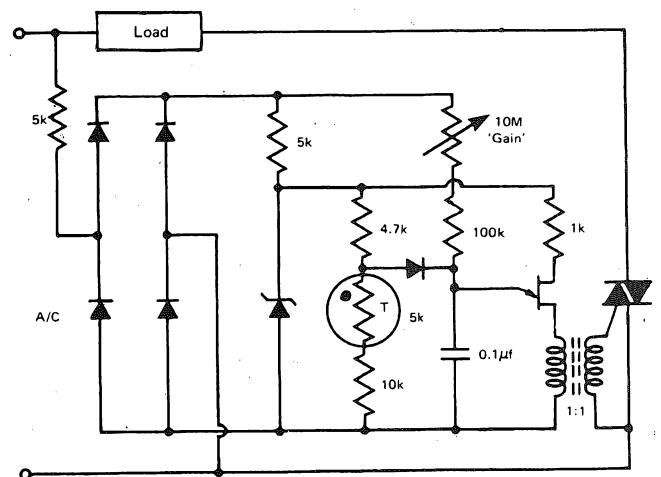


Fig. 26
This circuit provides proportional control of heating loads.

equipment.

The radiated energy can be reduced by including an rf choke in series with the power line (shown as L1 in circuits accompanying this article). A suitable choke can be made by winding 40 to 50 turns of enamelled copper wire around a piece of ferrite aerial rod. This will provide an inductance of approximately 100 μH . For greater suppression it will be necessary to increase the inductance to as much as 5 μH , and this necessitates a full-size choke.

The limitation of this type of circuit is that the rf choke has to carry the full load current, and once this exceeds a kilowatt or two, the physical size of the choke becomes excessive. The larger the load, the more difficult and expensive it becomes to filter out the rf components. (The problem is not as acute for inductive loads such as electric motors where the load itself attenuates the rfi).

Phase control also introduces another problem — namely the load power factor. This is adversely affected and is something that power supply authorities regard with less than equanimity.

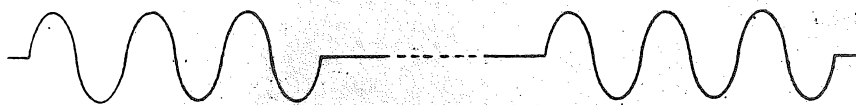


Fig 27
Zero voltage switching waveform — power is controlled by varying the ratio of complete half-cycles 'on' to complete half-cycles 'off'.

Zero voltage switching

Virtually total freedom from rfi can be achieved by a triac control system known as zero voltage switching.

This system differs from phase control in that line voltage is switched 'on' as well as 'off' only at zero crossing points of the sinusoidal waveform. The power applied to the load is controlled by varying the ratio of time that power is applied, to the time that power is switched off. For example, if only a small amount of power is required, the triac is caused to pass just a few complete half-cycles, then remain switched off for a further 30 or 40 complete half-cycles before repeating the sequence (Fig. 27).

It will be apparent that the zero voltage switching

system can only be used to control loads that have considerable inertia, such as heating elements which by their nature will average out the bursts of energy. Zero voltage switching cannot be used for controlling lighting or motor speed.

Load power factor is not affected in any way by zero voltage switching.

Until recently the drawback to zero voltage switching was that the technique required a large number of components.

Fortunately the functions necessary for zero voltage switching have been made available in an integrated circuit — and, by using this specialized ic, zero voltage switching circuits can be built that are outwardly no more complex than their phase control equivalents.

PRACTICAL GUIDE TO TRIACS PART 3

Our final article in this three part series explains how triacs are used in zero-voltage switching circuits — and includes circuits for the control of large heating loads.

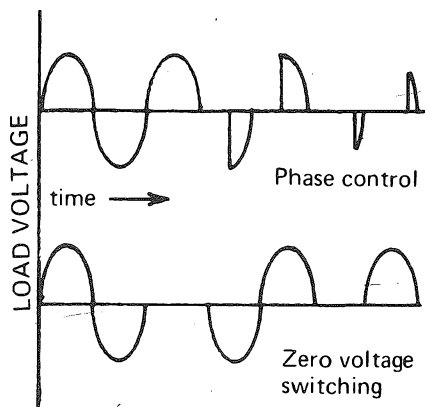


Fig. 28

Zero voltage switching is a method of varying the power applied to a load by switching line voltage on and off only at the zero crossing points of the sinusoidal waveform.

The technique virtually eliminates the problems of rfi associated with phase control of large resistive loads.

The difference between phase control and zero voltage switching is illustrated in Fig. 28, where the upper waveform shows phase control, whilst the lower waveform shows zero voltage switching.

Zero voltage switching can only be used for applications in which the controlled load is capable of averaging bursts of complete half-cycles. Thus the response time of the load must be long compared with the period of the switching cycle, for it is quite possible that at low power settings, short bursts of say, twenty to thirty half-cycles may be applied at ten second intervals. Thus the method is acceptable for the control of loads such as heating elements, but out of the question for light dimming.

Half-wave control

A very simple yet effective zero voltage switching circuit is shown in Fig. 29. This circuit provides half-wave control only, but is very satisfactory for commercial applications where the heating elements can be designed to suit.

The circuit is extremely stable and unaffected by quite large variations in line voltage and ambient temperature. The response time depends upon the characteristics of the thermistor which is used — times of one to two seconds are typical. The sensing differential is around 0.2 C at normal ambient temperature.

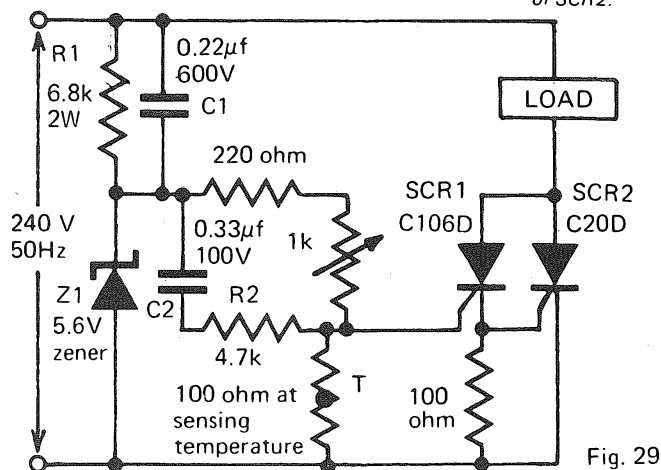


Fig. 29. Simple half-wave zero-voltage switching circuit, maximum load is determined by choice of SCR2.

Fig. 29

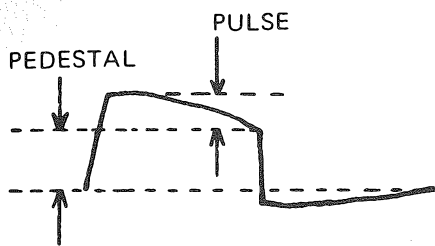
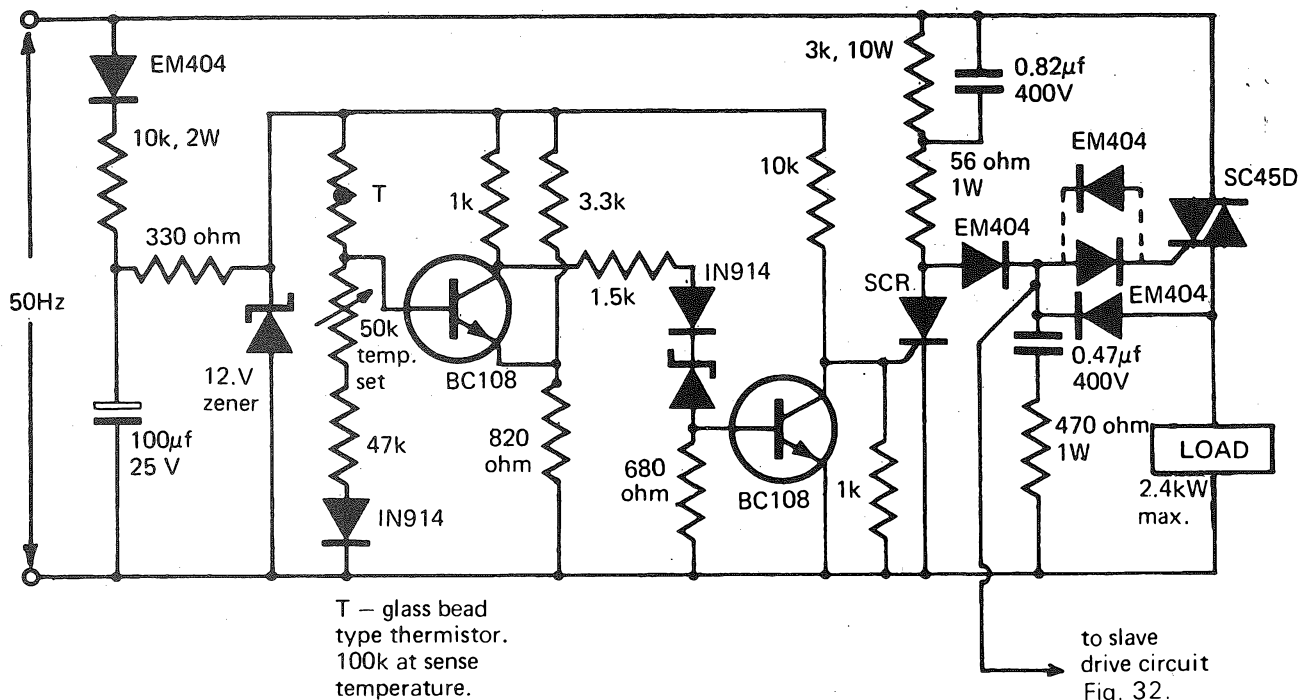


Fig. 30. Basic waveform associated with circuit of Fig. 29.

Fig. 30

Fig. 31. This circuit can control loads of a size determined by the rating of the Triac. The slave circuit shown in Fig. 32 can be driven by the output of this circuit if required.



The zener diode ZD1 forms a voltage pedestal of 5.6 volts nominal amplitude by clipping the incoming positive half-cycle of mains voltage. This pedestal is differentiated by R2, C2 and associated resistors to form a pedestal of reduced amplitude with a pulse superimposed on top of the pedestal (Fig. 30). This waveform is applied to the gate of SCR1. The capacitor C1, which is connected in parallel with R1, provides a leading phase shift to the pedestal so that SCR1 is triggered into conduction by the peak of the positive decaying pulse which is superimposed on the pedestal. It does this at the beginning of the positive going half-cycle of line voltage appearing at the anodes of both SCRs.

The thermistor controls the amplitude of the pedestal and this provides a semi-proportional control with a small temperature differential.

The lock-in configuration of SCR1 and SCR2 reduces the effects of ambient temperature variations. The cost of this circuit is very low compared to a phase control circuit of the same power handling capacity as no rfi components are required.

Full-wave control

The circuit shown in Fig. 31 will provide full-wave control of heating loads of almost any size. The triggering circuit will drive triacs of any size from 1 amp up to 125 amps. In addition almost any number of addition-

al triacs can be slave driven by the main triggering circuit. (The slave triggering circuit is shown in Fig. 32.)

The differential of this circuit is approximately $\pm 1/6^\circ\text{C}$. This circuit has a semi-proportional action, and is suitable for applications where large amounts of power have to be controlled accurately and at low cost.

Integrated circuit control

A number of companies have produced integrated circuits specifically for zero voltage switching applications. These ICs permit circuit and operational techniques which would otherwise require an unrealistic quantity of discrete components.

One typical zero voltage switching integrated circuit is the μA742 from Fairchild Semiconductor, 230 High Street, Potters Bar, Herts. When used with only a few passive components, plus a thermistor, this IC develops its own power supply voltage, differentially detects sensor imbalance, detects the zero voltage crossing point and produces the required triac triggering pulse.

Zero voltage switching integrated circuits of this type may be used in very sophisticated applications. One example of this is the 'fail-safe' control of heating (and refrigeration) loads.

In many control systems the failure (short or open circuit) of a thermistor or resistance thermometer may be almost literally catastrophic. For instance a heating

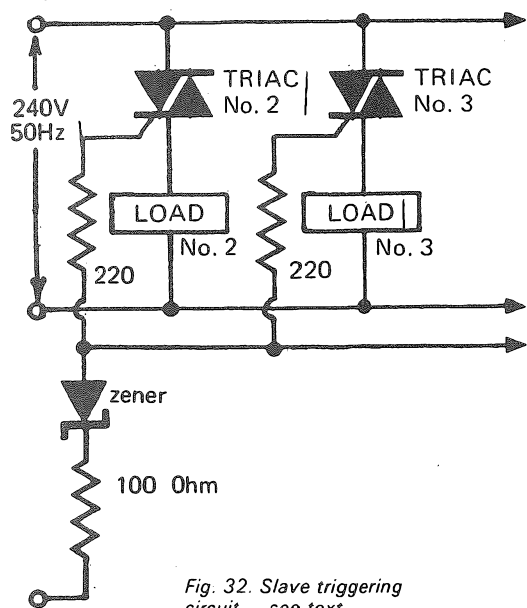


Fig. 32. Slave triggering circuit — see text.

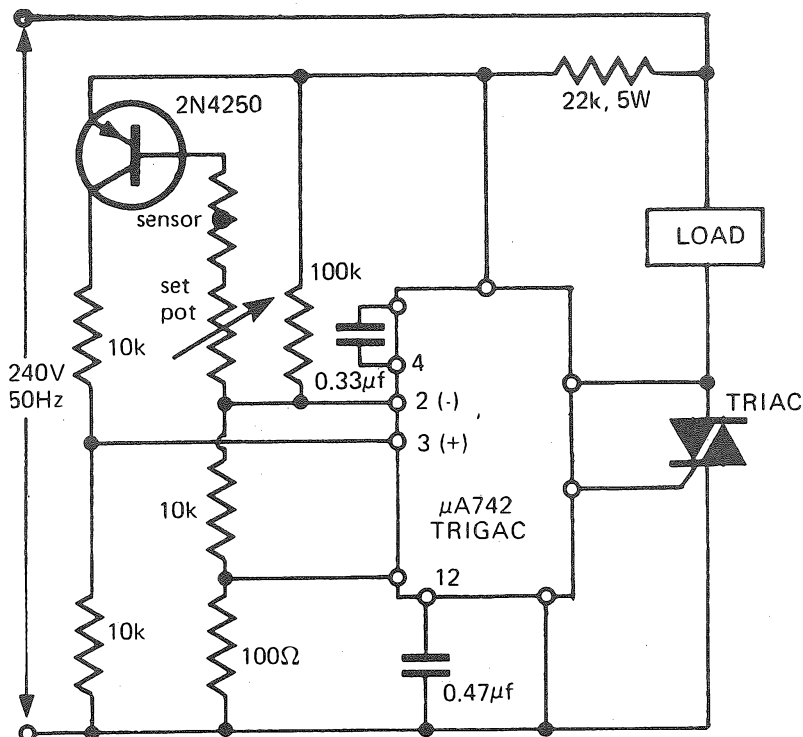


Fig. 33. This circuit using a Fairchild TRIGAC IC senses and protects heating loads against open circuit thermistors.

control with a negative temperature coefficient thermistor would interpret shorted thermistor leads as a very high sensed temperature and would react by cutting off the power. However if the same thermistor were to go open circuit (and this is by no means unknown) the control system would see this as a low temperature and full power would be applied continuously to the load. In this case an 'open circuit' thermistor detector is required to protect the system against this condition.

A circuit that will provide this protection is shown in Fig. 33. A slightly modified version of this circuit can be used to protect against 'short circuit' failure of positive temperature coefficient thermistors.

Zero-voltage switching may be used in conjunction with proportional control circuitry for both single and three phase loads. These types of systems are however necessarily complex and outside the scope of an article such as this. Full details of such circuits can be obtained from the manufacturers of zero voltage switching ICs.

Triacs — fault finding

As far as the practical electrician is concerned, triacs are relatively simple minded devices. They either work or they don't — there is rarely a halfway stage.

By far the greatest cause of failure is overloading. It cannot be stressed too strongly that a triac, like most semi-conductor devices, is destroyed instantaneously by

a short circuit placed across its output.

An almost infallible indication of an overloaded triac is that gate control will have been lost, and the triac is 'on' all the time. Checking this is quite simple, just unsolder all the connections to the triac's gate, and, if the triac is still switched 'on' then the unit has lost gate control.

Before replacing the triac, check with an ohmmeter to make sure that the load is not shorted out. When the circuit is again in operation, monitor the line current for a time to check that it is within the triac's designed rating. In particular remember that large incandescent bulbs have a very heavy inrush current and this may exceed the triac's short term overload capability. (This is generally five to ten times full rated load for one half-cycle of the input waveform.)

The other extreme of failure is when a triac circuit will not trigger at all. This is generally a failure of the triggering circuit rather than the triac.

The first obvious check is to establish that there is in fact mains voltage across the triac. Assuming that voltage is there, a very simple check is to connect a resistor (150 ohms to 1k) between the case and the gate. This should trigger the triac into full conduction.

If the triac is triggered into conduction, then the fault lies in the triggering circuits; if the triac does not trigger, then the triac is faulty.

Finally, don't test triacs with a Megger.

Touch Switch

For turning on equipment sensitive to vibration — photographic enlargers for example — or even just for a doorbell, this circuit is a boon



Photon Phone

The PHOTON is a light-talker: a communications system which requires no licence

THIS IS NOT a mere flash-light talker. It is a communications system which requires no licence. It is a light-talker. It is a communications system which requires no licence. It is a light-talker. It is a communications system which requires no licence.

USING IT
This circuit is a light-talker. It is a communications system which requires no licence. It is a light-talker. It is a communications system which requires no licence.

Starburst

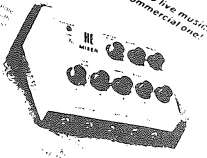
Expand your universe with HE's pulsating photonic display

LOOKING FOR A NOVEL DISPLAY? HERE IT IS!

FIND THE LIGHT
This circuit is a light-talker. It is a communications system which requires no licence. It is a light-talker. It is a communications system which requires no licence.

Audio Mixer

This project should prove invaluable to anyone who's into live music. It's unless they've just spent five times as much on a commercial one!



Ultrasonic Switch

Multi Option Siren

COMPLETE C
This circuit is a light-talker. It is a communications system which requires no licence. It is a light-talker. It is a communications system which requires no licence.

White Noise Effects Unit

Simulate the soothing sounds of wind, waves, and surf on the beach, or the roar of jet aircraft and steam trains, with this HE White Noise Effects Unit.

Constant Volume Amplifier

The HE Constant Volume Amplifier

Shark!

H.E. Shark game brings you absolutely no water!

Graphic Equaliser

Getting Started

Before you jump in at the deep end, here are some words of advice for beginners

Digital Clock

Pass some time with this time project! Using only the minimum components you can construct multipurpose digital clock

PROJECTS SPECIAL VOL 1

£1.25 FROM HOBBY ELECTRONICS

For your book of HE Project Reprints send £1.25 + 25p (P&P) to Sales Office (Specials), Hobby Electronics, 145 Charing Cross Road, London, WC2H 0EE.

Transistor Gain Tester

Tester is somewhat unusual in that it does not use a visual indication of the transistor's gain

Linear Scale Ohmmeter

HE's latest accurate piece of test gear. An inexpensive unit that gives rapid and accurate readings of resistance from a few tens of ohms to one megohm.

Tantrum

Want a first rate 25 watt stereo amplifier? Look no further project, turn up the volume and watch the neighbours live

Push-Button Dice

We think it's a good idea to have a few dice on hand for those occasions when you need them

Envelope Generator

Here's a project that you can use to simulate the sound of an envelope generator for your synthesizer

Digibell Project

Using door bells with the digibell project will never be ignored

LED Tachometer

The HE Tachometer uses 21 LEDs to give a solid-state analogue RPM display. It's an ideal project for the motor or motor-cycling enthusiast

Drill Speed Controller

An easy-to-build gadget that lets you vary the speed of your drill and other domestic appliances that are powered by a.c. units

Parking Meter

Avoid expensive parking fines with this pocket-sized warning device

Analogue Audio Frequency Meter

Get in tune with the modern age... economical and accurate made to measure

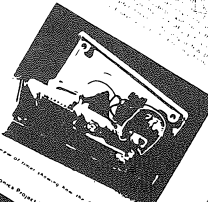
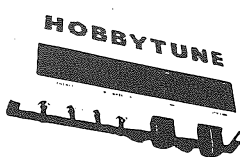
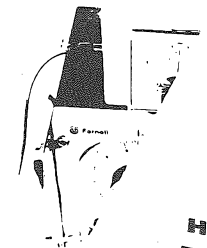
Hobbytune

Build a full feature monophonic organ for less than ten pounds

Photographic Timer

This project should appeal to the photo enthusiast. Use it to do your enlarger timing

THE LIGHT FANTASTIC



POWER SAVING

CONSTRUCTION

POWER SUPPLIES

One more from Tim Orr. This time he takes us through a series of different methods for powering up circuits. On the way he explains the theory behind each.

The job of producing stable regulated power rails has been much simplified by the introduction (about seven years ago), of three terminal fixed voltage regulators. These devices can make the power supply design problem relatively simple, but even so the designer must be fully aware of a lot of other important details that can cause poor results. Firstly, consider a simple unregulated power supply, fig. 1.

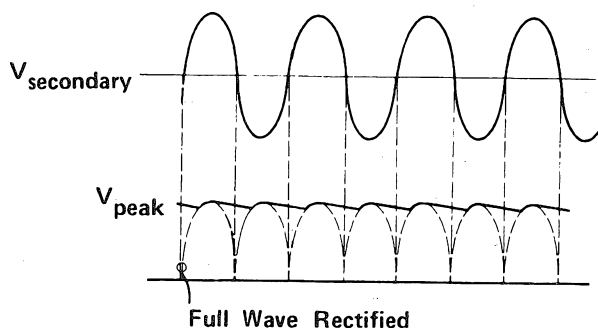
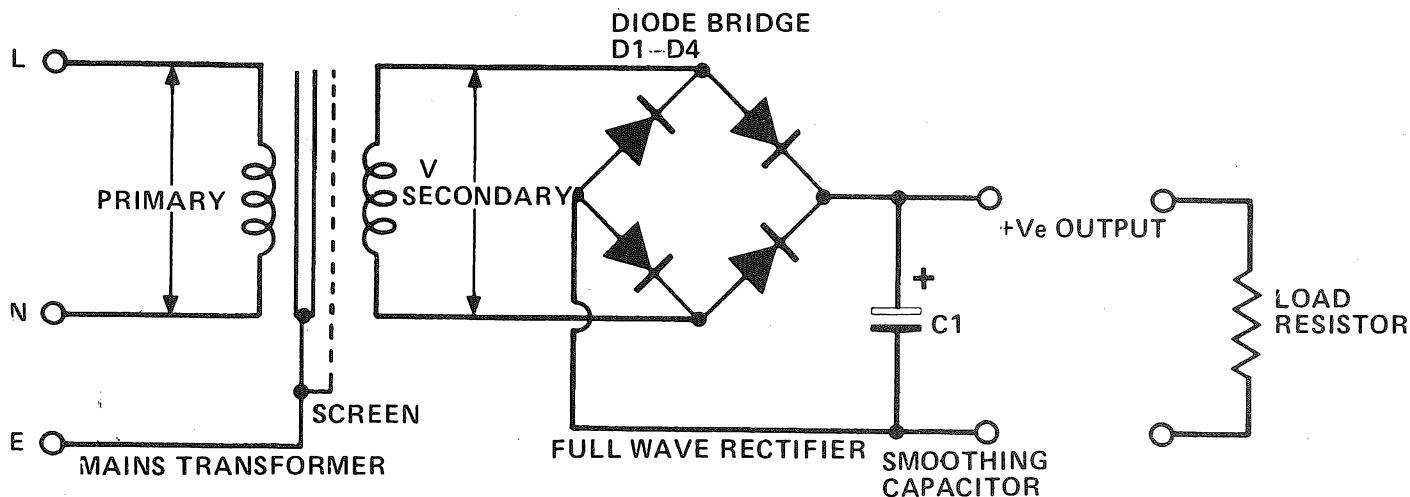


Figure 1. Below: an unregulated power supply. Above: The output (with a load resistor).



This piece of hardware has three sections, a step down, isolating transformer, a diode bridge and a smoothing capacitor. The transformer is driven from the mains, the voltage of which varies depending on where you live (it's 250V_{RMS} in Fulham). Some transformers have got a copper screen which isolates the primary winding from the secondary windings. For the purpose of safety, this should be connected to earth.

Also, for maximum safety, connect the 220/240/250 tapping to mains LIVE. Another type of mains transformer uses what is known as a split bobbin, the primary is wound on one bobbin, the secondary on another. Thus the two windings are inherently physically isolated, and so no safety screen is included. These two transformer types are generally constructed on what is known as an 'E' core; take one to bits and you will find that it is

The function of a mains isolating transformer is to physically separate the user end of a piece of equipment from the 'potentially' (!) lethal mains voltage. The transformer also provides a suitable voltage which can be rectified and smoothed and connected to a voltage regulator. This is the secondary voltage of a transformer and it is measured in V_{RMS} at a particular loading.

That is, if the transformer is rated at 15V_{RMS} at 10VA, then the output voltage will be 15V_{RMS} when the load upon the transformer secondary is 10VA (10 watts).

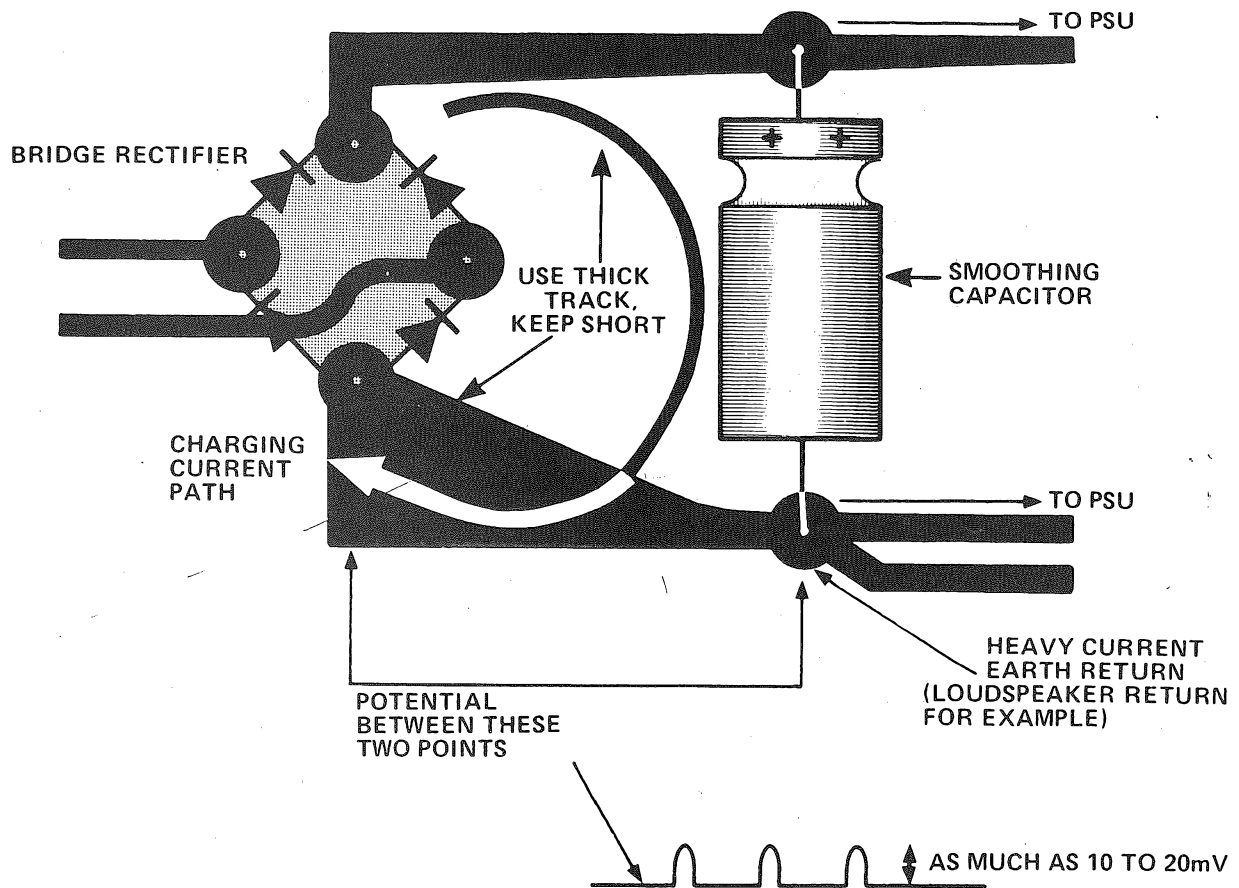
If the load is removed the output voltage will rise. The percentage change from load to no load is known as the TRANSFORMER REGULATION and is typically of the order of 20%.

To convert the V_{RMS} voltage to a DC voltage it must be multiplied by 1.4142. Thus a 15V_{RMS} (loaded) transformer secondary will generate 21V_{DC} when full wave rectified and smoothed, which will rise to 25V_{DC} when the load is removed (assuming 20% regulation see Fig. 1).

Thus care has to be taken when selecting a transformer such that the smoothing capacitor working voltage is not exceeded. Also, make certain that the polarity on this capacitor is correct, they can LITERALLY explode if wired up backwards!

constructed out of lots of 'E' shaped laminations. These 'E' laminations are butted into 'I' laminations, and clamped together. This butting together of the laminations can cause magnetic field problems. The wider the gap between the 'E' and 'I' laminations, the larger the magnetic field around the transformer.

The magnetic field generates a significant amount of induced hum in nearby electronics, this can be overcome by using a low leakage toroidal transformer which is constructed from circular laminations. The primary and secondary windings are wound through the centre of the toroid (see if you can imagine how). The toroidal transformer, by virtue of its 'continuous' laminations results in a low stray field and a low profile design, making it ideally suited for audio amplifier applications.



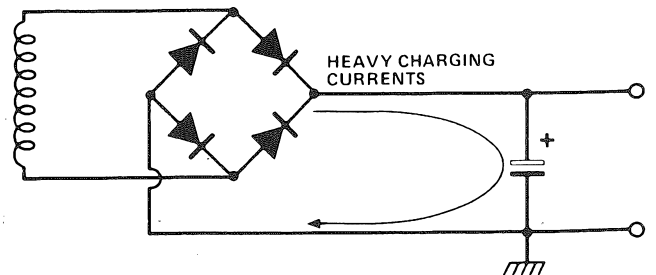
When a load is placed upon the power supply shown here, the output voltage appears as a DC voltage on top of which is a ripple voltage. This can be thought of as two separate periods, a charge period where the capacitor is charged up by the power supply and a discharge period where the load discharges the capacitor.

This charging and discharging generates a ripple voltage which has a period of 10 ms (100 Hz). A load current of 100 mA, and a 1000u capacitor will result in a ripple voltage (Vpp) of about 0V7.

As a rule of thumb I usually allow 1 to 1V5 maximum ripple if a voltage regulator is being used. This will generally result in an output ripple of less than 1 mV. If this ripple were to be obtained by just using a larger capacitor, then a 700,000u capacitor would be required!

Generally the discharge period is much longer than the charge period. This means that the transformer is only supplying power for short periods, in fact during the charge period. During these periods the smoothing capacitor is rapidly charged, and it is quite common for these current surges to exceed several amps. This can cause mains BUZZ problems when laying out printed circuit board designs for power supplies.

The correct layout is shown above the circuit. If the current surge is 1 A and the track resistance is 20 milliohms then the voltage developed will be 20 mVpp.



Voltage regulators

A voltage regulator takes a varying unregulated input voltage and produces a fixed regulated output voltage. There is a wide range of fixed voltage three terminal regulators to choose from, with a choice of maximum current handling, output voltage and positive or negative operation. The data sheets for these devices contain lots of seemingly complex pieces of information and so a glossary of terms is now included.

Ripple Rejection

The ratio of the ripple voltage at the regulator input to that at the output, generally expressed in dB. Typically of the order of 60 dB (1000 to 1), that is 1 Vpp of ripple at the input ends up as 1 mVpp at the output.

Temperature Coefficient

The output voltage change for a change in regulator temperature, expressed in mV/°C.

Input Voltage range

The range of voltages over which the regulator will function normally. For example, a 12V regulator may work from 14V5 to 30V. At 14V5 the regulator will 'drop out' and lose its regulation. Regulators generally need 2 to 2V5 in excess of their output voltage. At 30V the regulator will go 'pop' (time to buy a new one).

Output voltage

The voltage at the output terminal with respect to ground. Generally within $\pm 5\%$ of stated value.

Line Regulation

The ratio of the change in the output voltage caused by a change in the input voltage, typically of the order of 0.2%.

Load Regulation

The output voltage change for a specific change in output load current.

Short Circuit Current

The output current when the output is shorted to ground.

Output Noise Voltage

The RMS noise voltage measured at the regulator's output, not including any ripple.

Power Dissipation

The maximum power that the regulator can safely generate on a particular heatsink.

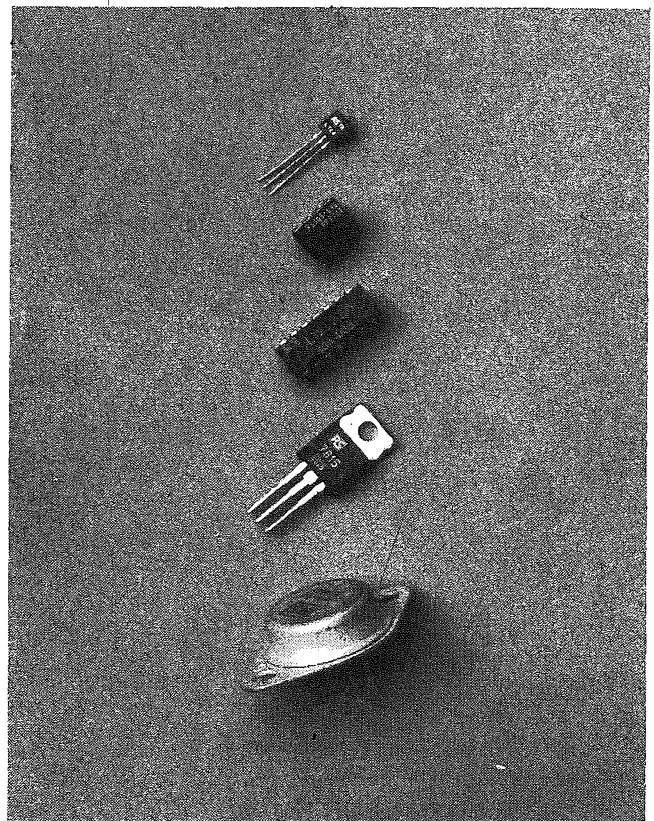
As a rule of thumb the regulator case should not exceed about 80° C (which is hot to touch). However, always run the device at as low a temperature as possible. It is thermal ageing that eventually kills electronic devices and for higher temperatures the ageing process is disproportionately faster.

Some applications of voltage regulators are given below.

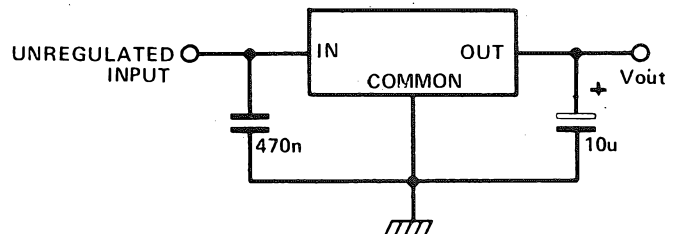
The table below relates the secondary voltage of a transformer to the peak voltage at rated load and the off load voltage, which will be considerably higher.

TABLE ONE

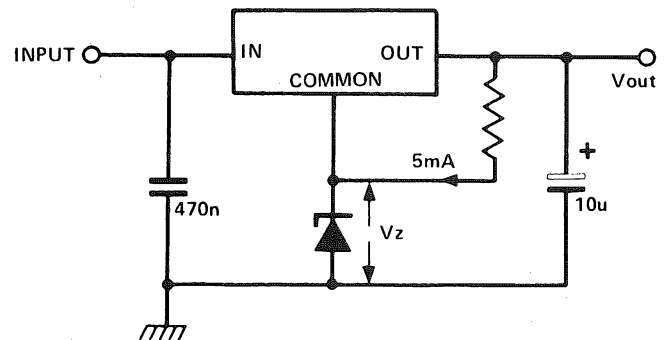
V secondary at rated load	V peak at rated load	V peak off load transformer regulation 20%
5 VRMS	7V07	8V48
6 VRMS	8V48	10V18
9 VRMS	12V72	15V26
10 VRMS	14V14	16V97
12 VRMS	16V97	20V36
15 VRMS	21V21	25V45
20 VRMS	28V28	33V93
25 VRMS	35V35	42V42
30 VRMS	42V43	50V92
35 VRMS	49V50	59V40
40 VRMS	56V57	67V88



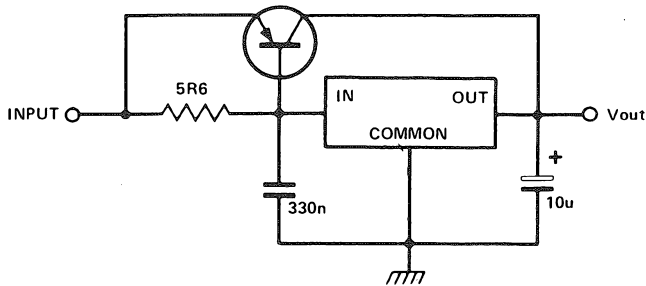
From top to bottom: The 78L12 (TO92 case), 12V 100mA; the 7727 (DIL), fixed $\pm 15V$ at 50mA per rail; the L723 (DIL), adjustable +ve or -ve at 100mA; the 7815 (plastic package), 15V at 1A; and the 78H05 (TO3), 5V at 5A.



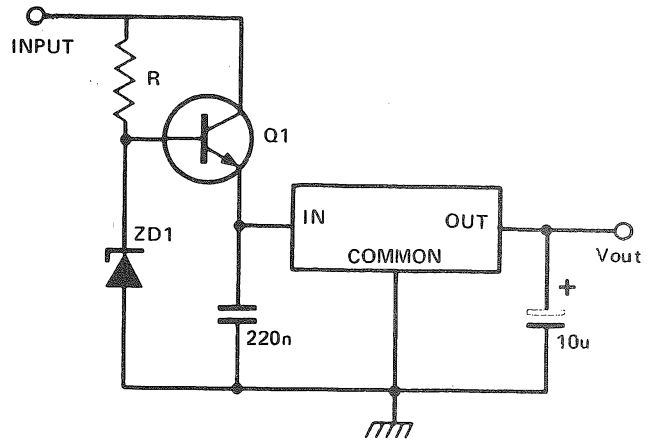
This circuit shows a conventional arrangement of a three terminal device. It is advisable to use a decoupling capacitor connected close to the input terminals. This prevents high frequency instability. If this capacitor is left out then regulation can sometimes be greatly reduced. The decoupling capacitor on the output helps reduce the impedance at high frequencies, where the regulator loses its performance. For best results use a tantalum capacitor.



The output voltage of a regulator can be increased by applying a voltage to the common terminal. This can be done by using a zener diode.



The output current can be increased by using a bypass transistor. When the current flowing through the voltage regulator exceeds 100mA (the voltage across the 5R6 resistor being 560mV), the bypass transistor begins to turn on. This transistor takes all currents in excess of 100mA and yet the output still remains regulated. However a few extra components are needed to get current limiting in the transistor path.

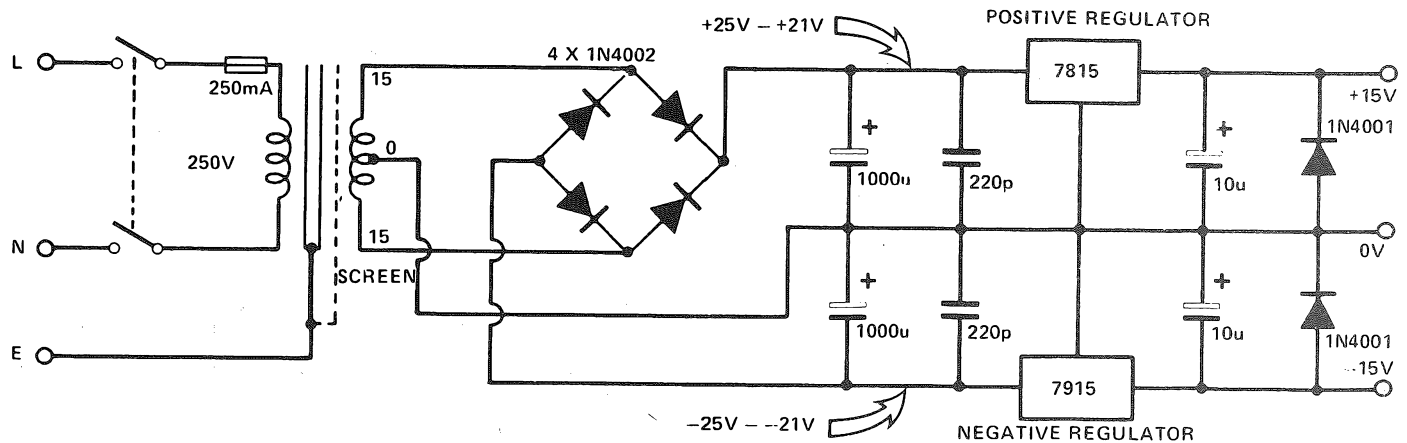


A high voltage unregulated supply can cause problems when using regulators. It may at times exceed the maximum voltage rating of the regulator. A simple voltage regulator ZD1 and Q1 can be used to overcome this problem. ZD1 should be chosen so that it is about 6V greater than the regulator output voltage. This technique has the added advantage that the power dissipated in the regulator is less (the rest being dissipated in Q1), and the regulator is presented with a semiregulated voltage, so the output will have less ripple.

Dual Power Supply

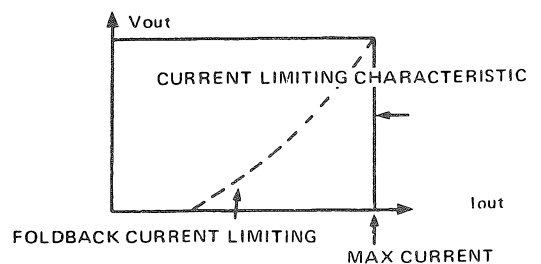
The circuit shows a complete regulated dual power supply. The unregulated rails are obtained from a split secondary transformer, a bridge rectifier and two smoothing capacitors. A positive and a negative regulator have been used to generate the + and - rails. These regulators should be mounted on heat sinks

and they should be insulated. The pin out of the negative regulator is different to that of the positive regulator. The two diodes at the output prevent latching up situations (on load) whereby one side starts up faster than the other and forcibly reverse biases it, preventing it from operating.



Tracking Regulator

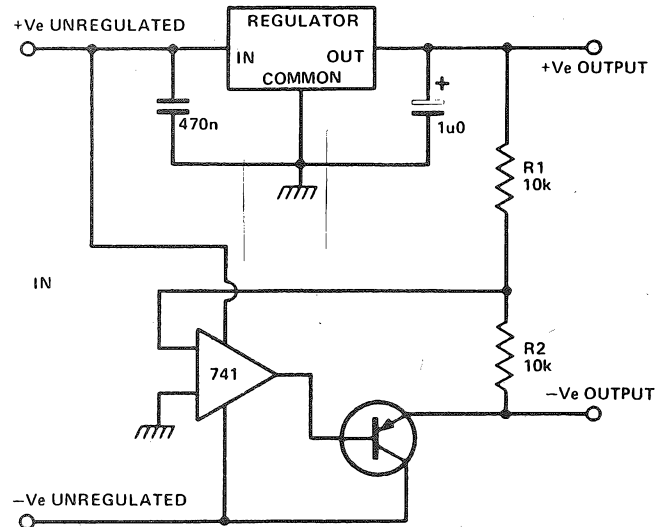
Instead of using a negative voltage regulator to obtain the negative rail, an op amp and a power transistor can be used. The resistor ratio, R1, R2 determines the negative rail voltage. The negative rail is not, however, current limited. The internal current limiting of the regulator is shown. When the load current exceeds the current limit, the output voltage drops to almost 0V. This makes the regulator short circuit protected. Another type of current protection is known as 'FOLD BACK' current limiting (shown dotted). This serves to reduce the short circuit current. These devices protect the power supply from abuse. Another type of protection device is the overvoltage clamp, which



protects the 'non-power supply electronics' from an increase in the power supply voltage. These are two terminal heavy current devices which are placed across the power supply. When the supply voltage exceeds a certain level a thyristor is triggered on and clamps the rail to ground. This is intended to pop a fuse and so disconnect the faulty power supply (which is better than replacing a £1,000 worth of ICs).

$$-ve\ output = -(+ve\ output \times R2/R1)$$

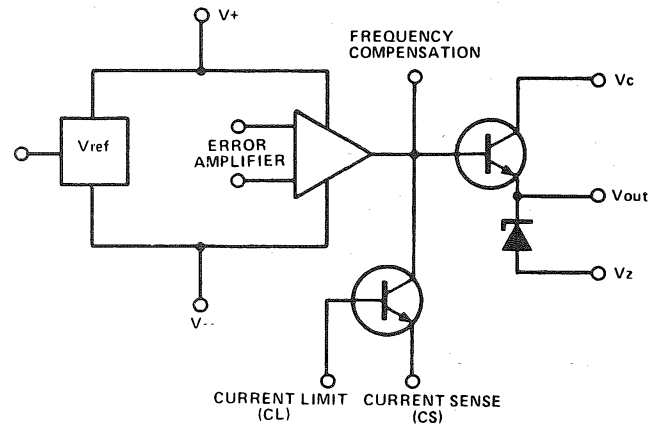
With foldback the short circuit power dissipated in the regulator is less than that with current limiting.



723 Voltage Regulator

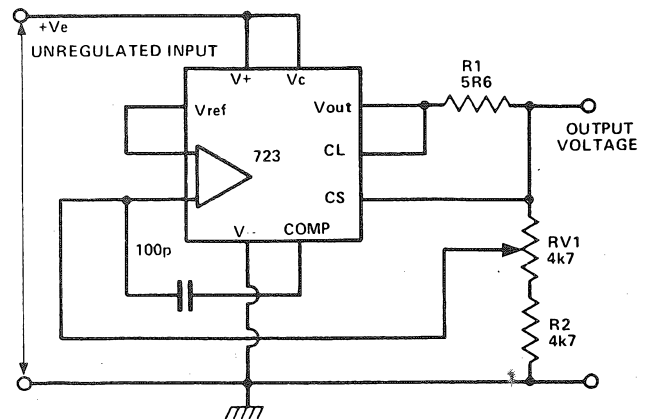
The 723 is an industry 'standard' device. Many manufacturers produce it and the device itself is versatile. It comes in a 10 pin TO5 can or a 14 pin DIL pack. The device contains a precision voltage reference, with a temperature coefficient of 50ppm/°C, an error amplifier, an internal transistor capable of handling 100 mA and a current limiting mechanism. By using a few external resistors, a capacitor and maybe an external power transistor, a wide variety of regulator designs can be realised.

Left is shown the block diagram of the 723 regulator. As pinouts vary depending upon package, no pin numbers are shown.



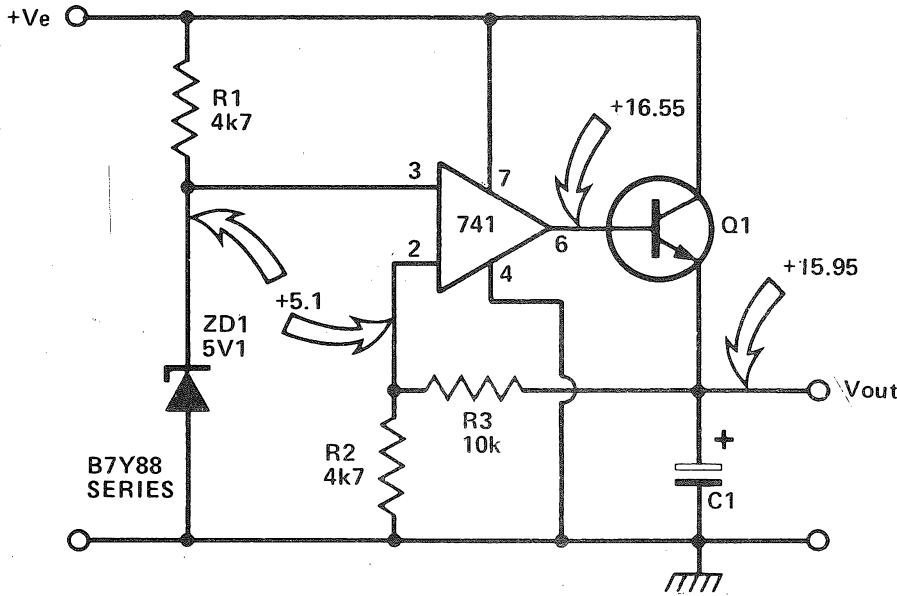
Adjustable Positive Voltage Regulator

By using a variable feedback path (RV1), a variable regulated output voltage can be generated. The voltage reference is connected to the non-inverting input of the error amplifier and the output voltage (via RV1), to the inverting input. The error amplifier drives the output transistor and hence the output voltage is controlled by the feedback voltage from VR1. A 100pf capacitor is used to stabilise the device. R1 is used as a current limit control. When the current through R1 (the load current), exceeds 100 mA a voltage of 560 mV is set up across it. This is just about sufficient to turn on the current limiting transistor which in turn shorts out the regulating transistor, causing the output voltage to collapse towards 0V.



Regulated Power Supply

Sometimes it is necessary to make a simple power supply using discrete components when a non-standard voltage is required.



Left: Circuit diagram of discrete component PSU. Voltage measurements are taken with high impedance voltmeter.

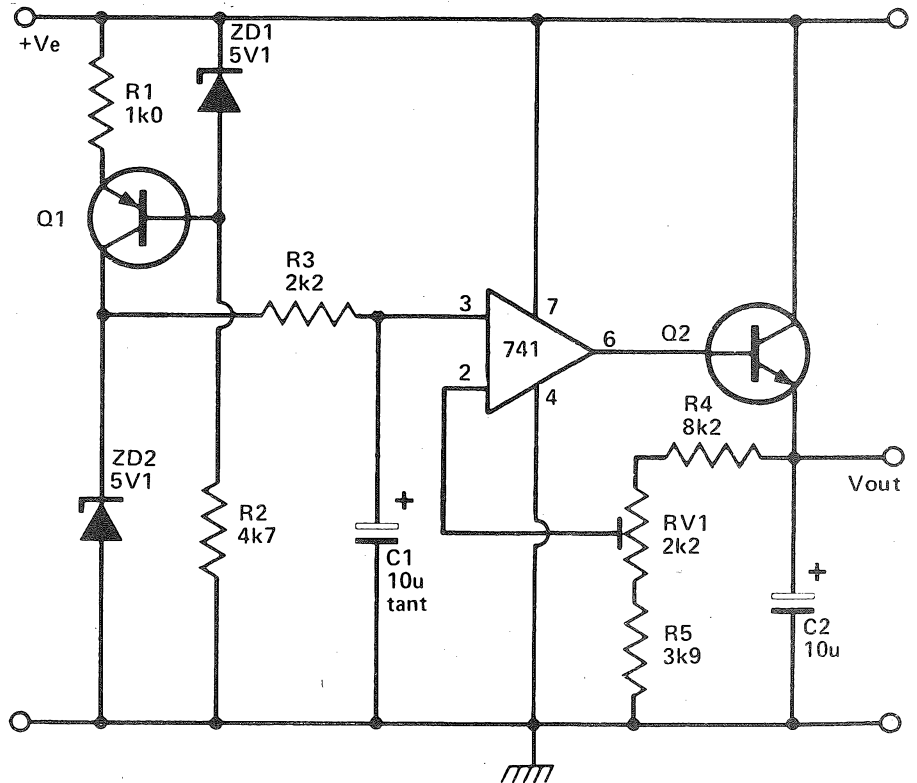
The circuit shown uses all the basic elements of a voltage regulator, that is, a reference voltage ZD1, an error amplifier and a series control transistor Q1. The zener diode, ZD1 sets up a reference voltage of 5V1. This diode has a temperature coefficient of $-1.2\text{mV}/^\circ\text{C}$ (a 5V6 zener is best at $-0.2\text{mV}/^\circ\text{C}$). The resistor ratio of R3 and R2 sets the output voltage and the op amp provides the error correction (the regulation).

C1 is used to reduce the output impedance at high frequencies. The zener diode has a slope resistance of $76R$, and so any fluctuations in the unregulated rail will be attenuated by the ratio of $76/4700$, or 0.016 (since R1 is $4k7$). Therefore a 1Vpp ripple will end up as 16mVpp , but will be multiplied by the gain of the R3, R2 network to nearly 50mV .

Improved Regulated power supply

This power supply has various improvements over that shown above. The reference zener ZD2 is run at almost constant current by the R1, R2, Q1, ZD1 network. This makes ZD2 much less sensitive to ripple and unregulated supply fluctuations. The filter R3, C1 (7 Hz low pass), further reduces any ripple voltage and noise from the zener diode. The preset RV1 allows the output voltage to be varied.

If a precision power supply is required then a precision voltage reference should be used. These can be obtained with temperature coefficients as low as $10\text{ppm}/^\circ\text{C}$. When using this level of stability, high stability resistors ($\text{TC}=10\text{ppm}/^\circ\text{C}$), and a low drift op amp should be used. Also, to reduce mains carried interference (mainly sharp clicks due to electric motors and thyristors turning on), a mains filter should be used. This is a passive inductor-capacitor low pass filter network which attenuates high frequency spikes and clicks.



A PRACTICAL GUIDE TO ZENER DIODES

Like the Establishment, but of more obvious value, zener diodes remain constant and unperturbed by external change.

Zener diodes — named after Dr. Carl Zener — are used primarily as voltage references. They are devices that maintain an almost constant voltage across them despite various changes in circuit conditions.

Externally, zener diodes look much the same as conventional diodes and are manufactured in axial lead, and stud mounting forms; furthermore zener diodes are capable of rectifying alternating current into pulsating direct current as are their conventional counterparts.

But unlike conventional diodes, zener diodes are deliberately intended to be used with the anode connected to a negative potential and the cathode connected to a positive potential.

When connected in this manner, zener diodes have a very high resistance below a certain, critical, voltage (called, appropriately, the zener voltage).

But if this critical voltage is exceeded the dynamic resistance of zeners drops to a very low level. And in this region, essentially constant voltages will be maintained across the zeners and these constant voltages will be maintained despite quite large changes in the applied currents.

This characteristic is illustrated graphically in Fig. 1 from which it may be seen that beyond the zener voltage the reverse voltage remains practically constant despite changes in reverse current.

Because of this characteristic, zener diodes may be used to provide a constant voltage drop, or reference voltage, across their internal resistance.

Zener diodes are manufactured in a number of wattage ratings and with zener voltages ranging from 2.7 volts to 200 volts. (In practice, zener diodes with ratings exceeding 30 volts are rarely used.)

The Basic Circuit

A basic voltage regulator circuit, using one resistor and one (idealized) zener diode, is shown in Fig. 2. In the example shown, the zener diode is rated at 5.6 volts and the applied voltage V_{in} is 8.0 volts. With no load applied to the output terminals 5.6 volts will be dropped across the zener diode and the remaining 2.4 volts will be dropped across resistor R_s .

If the input voltage is now changed say, from 8.0 volts to 9.0 volts, then the voltage drop across the zener will still remain at its nominal 5.6 volts but the voltage drop across R_s will increase, from 2.4 volts to 3.4 volts. The voltage drop across the (ideal) zener will remain constant.

In practice, the voltage will increase slightly due to the dynamic resistance of the zener. (The method of calculating the change in zener voltage is simply to multiply the dynamic resistance of the zener diode by the change in zener current.)

The resistor R_1 , in Fig. 2, represents an external load. If this load is connected across the circuit, then some of the current that was flowing through the zener will now pass through the load. Providing the current in R_s is greater than the load current some current will still pass through the zener and a substantially constant voltage will be maintained across the zener/load.

The series resistor R_s is selected so that the minimum current passing through the zener is not less than the level required for stable regulation.

This level begins just below the 'knee' of the reverse voltage/reverse current curve (as shown in Fig. 1). Curves such as this are available for all makes and types of zener diodes.

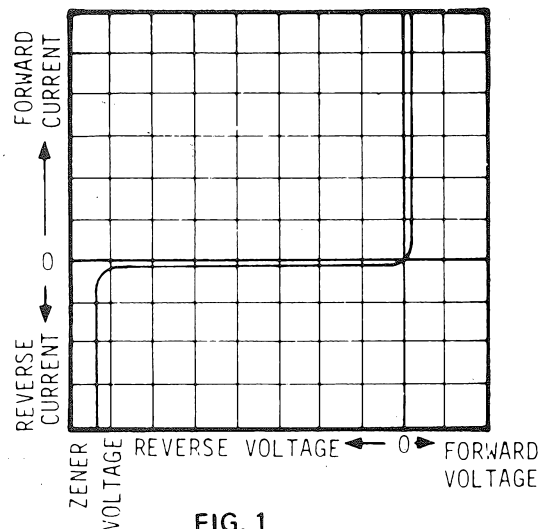


FIG. 1

It is also necessary to ensure that the value of R_s is such that current flow through the zener diode cannot exceed its specified power rating: which is the zener voltage multiplied by the zener current. In the circuit shown in Fig. 2, maximum current will flow through the zener diode when the load is disconnected.

The design procedure for this circuit is quite simple.

1. Specify the maximum and minimum load current (I_L), say 10 mA and 0 mA.
2. Specify the maximum supply voltage that is likely to occur (say, 12 volts), but ensure that the minimum supply voltage will always be at least 1.5 volts higher than the zener voltage of the diode to be used.
3. In the circuit shown in Fig. 2, the required output voltage, and hence the zener voltage, is 5.6 volts and the specified minimum zener current is 100 microamps. Thus the maximum zener current is 100 microamps plus 10 milliamperes — which is 10.1 milliamperes.
4. The series resistor R_s must conduct 10.1 mA at the

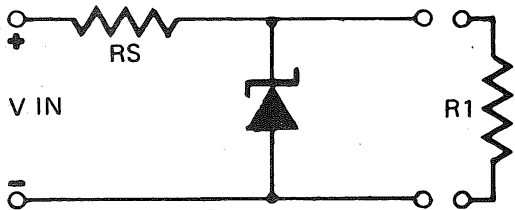


FIG. 2

lowest input supply voltage: and so allowing 1.5 volts minimum voltage drop across R_s (i.e. input voltage minus zener voltage):—

$$R_s = \frac{1.5}{10.1 \times 10^{-3}} = 148.5 \text{ ohms}$$

- The value of R_s is thus 148.5 ohms, and the nearest preferred value to this is 150 ohms.
- At the maximum supply voltage (12 volts), the voltage drop across R_s is $I_z R_s$, (I_z being zener current):—

$$\text{thus } I_z = \frac{(12 - 5.6) \text{ mA}}{150} = 42.6 \text{ mA}$$

- This is the maximum current that will flow through the zener at any time, i.e., maximum input voltage and zero external load. The power dissipated by the zener under these conditions is:—

$$= I_z V_z = 5.6 \times 42.6 = 238 \text{ mW.}$$

- Having calculated that the correct value for R_s is 150 ohms, and that the zener diode must be capable of dissipating 238 mW all that remains is to choose the correct type of 5.6 volt zener. This is, in fact, well within the capabilities of most small zener diodes which are generally rated at 400 mW.

A Portable Radio In Your Car

A practical example of the circuit described above is illustrated in Fig. 3. This shows how to run a transistor radio from your car's power supply.

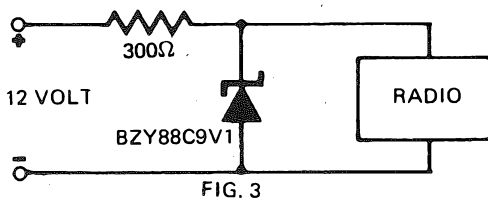


FIG. 3

The voltage required for the radio is 9 volts — and the nearest zener diode to this is 9.1 volts.

The maximum current drawn by the radio is measured at — say — 10mA, and, as the load does not vary greatly, this is also the minimum current. (As 10 mA is well above the 'knee' current of most small diodes, the addition of a minimum operating current is not required.)

The maximum supply voltage is around 14 volts (when the battery is fully charged).

Thus the series resistor R_s must conduct 10mA at the

lowest supply voltage (12 volts) and its value is calculated by:—

$$R_s = \frac{12 - 9.1}{10 \times 10^{-3}} = 290 \text{ ohms}$$

At the maximum supply voltage (14 volts) the voltage drop across R_s is $I_z R_s$:

$$I_z = \frac{(14 - 9.1)}{290} = 16.8 \text{ mA}$$

The maximum power dissipated in the zener is:—
 $9.1 \times 16.8 \text{ mA} = 152.6 \text{ mW}$

This is well within the capabilities of a standard 400 mW zener, such as a BZY 88 C9V1.

Temperature Drift

Although zener diodes are relatively immune to changes in supply voltage and load, they are to varying extents affected by changes in ambient and operating temperature. Fig. 4 shows a typical zener diode temperature coefficient curve, and as can be seen, the coefficient, although approaching 0.1% per degree Centigrade at the higher voltages, passes through zero at about 5 volts and then becomes negative for lower voltages, reaching $-0.4\%/^{\circ}\text{C}$ at approximately 3.5 volts.

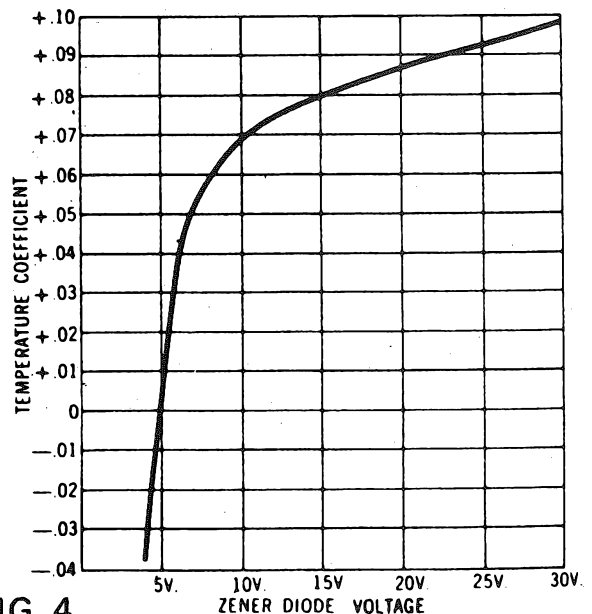
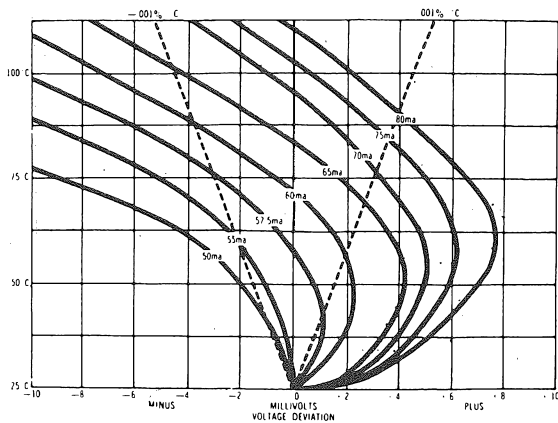


FIG. 4

The transition between a negative and a positive temperature coefficient is not well defined and a 5.6 volt zener may have either a positive or negative coefficient depending upon the current flow through it. But by careful control of the operating current it is possible to hold a standard 5.6 volt zener diode to a temperature coefficient of $\pm 0.001\%$ over a temperature range of $+25^{\circ}\text{C}$ to $+75^{\circ}\text{C}$ (Fig. 5 refers).

The only difficulty is to provide a constant current source for the zener diode. One method is to use a 10 volt zener diode to act as a regulator for the reference



diode. (Fig. 6). The current limiting resistor between the two diodes should have a negative temperature coefficient to compensate for the positive (+0.07%/°C) temperature coefficient of the 10 volt zener.

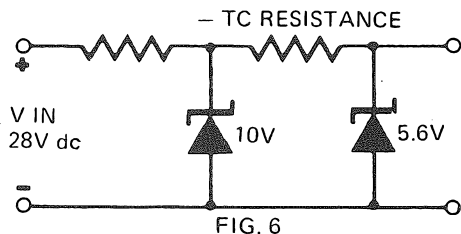


FIG. 6

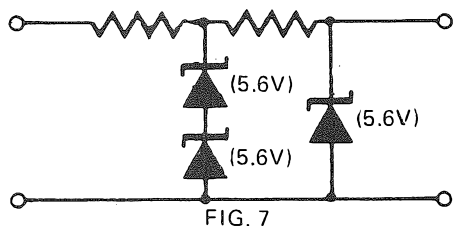


FIG. 7

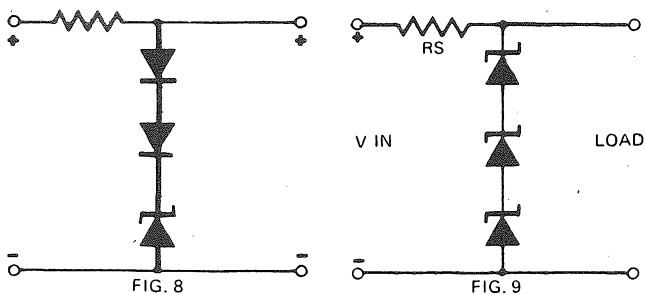


FIG. 8

FIG. 9

A second method of stabilizing the current to the zener diode is shown in Fig. 7. In this arrangement two 5.6 volt zeners, each having practically zero temperature coefficient, in turn stabilize the current through the voltage reference.

Another approach to compensation for temperature change is shown in Fig. 8 where the temperature coefficient of forward-biased silicon diodes is used to counteract the opposite temperature coefficient of the zener diode. This method may be used to compensate for voltage drift over quite wide ranges of both temperature and voltage, zener voltage stabilities of 0.01% are readily achieved.

As can be seen from Fig. 4, the temperature coefficient of zener diodes increases considerably at the higher zener voltages. An alternative method of temperature stabilizing to that shown in Fig. 8, is to use a

number of 5.6 volt zeners in series. (Fig. 9). The output voltage must, of course, be a multiple of 5.6 volts.

Temperature Sensing

The apparent disadvantage of a zener diode's temperature coefficient may be put to a useful purpose in the form of a temperature sensing device. Fig. 10 shows how a bridge consisting of two resistors and two similar zener diodes may be constructed so as to indicate a temperature level when one of the diodes is held at a reference temperature and the other is subject to the conditions to be monitored. The average small 10 volt zener has a temperature coefficient of +0.07%/°C — this corresponds to 7 millivolts per °C change. The sensing element will, therefore, indicate an imbalance of 0.7 volts when undergoing a 100°C temperature change.

Non-standard Voltages

Occasionally it is necessary to obtain a regulated voltage other than that obtainable from a single zener,

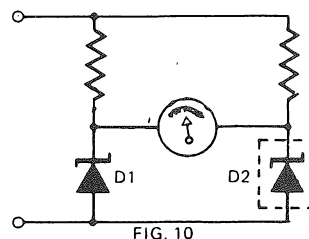


FIG. 10

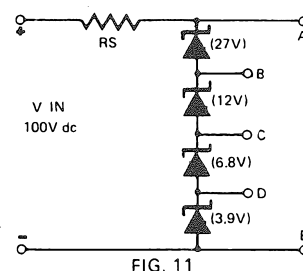


FIG. 11

this can be achieved by connecting a number of zener diodes in series (Fig. 9). The diodes need not have equal breakdown voltages since the arrangement is self-equalizing. However, the power handling ability of each diode should be the same. In addition, the current ranges should be similar or the loads so arranged to avoid damaging any of the diodes.

A group of zener diodes may be used as a voltage divider to obtain several regulated voltages simultaneously. (Fig. 11). This circuit may be used as a meter or scope calibrator. The four diodes shown in the circuit will supply 10 possible voltages: —

Voltage	Terminals
3.9	E — D
6.8	C — D
10.7	C — E
12.0	B — C
18.8	B — D
22.7	B — E
27.0	A — B
39.0	A — C
45.8	A — D
49.7	A — E

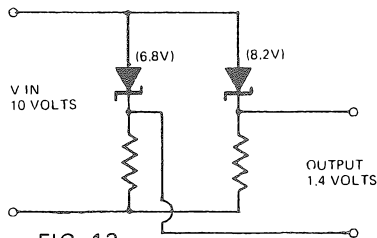


FIG. 12

It may also be necessary at times to provide a regulated voltage lower than the 2.7 volts minimum normally obtainable for a zener diode. Voltages lower than 2.7V may be obtained by using the *difference* in potential between a pair of zeners, (Fig. 12). The temperature compensation of this circuit is excellent, for both zener diodes tend to drift in the same direction, thus maintaining the difference voltage.

A PRACTICAL GUIDE TO ZENER DIODES PART 2

The zener diode is generally associated with dc applications, such as the control and regulation of dc power supplies. Most zeners are in fact used for this purpose but nevertheless they have many uses in ac, audio, rf, and ac control systems.

When supplied with alternating current and connected as a shunt regulator (Fig. 13), the zener diode will limit both the positive and negative halves of the ac cycle. The diode conducts almost immediately after the signal passes through zero and into the negative segment. On the positive half cycle the diode does not conduct until the applied voltage reaches the zener voltage (Fig. 14). The result is a non-symmetrical square wave. Asymmetry can be reduced by using high input voltages, but can never be completely eliminated unless two shunt connected diodes are employed in a back-to-back configuration. (Fig. 15).

The zener diode configuration shown in Fig. 15 is often used to provide stabilized filament supply voltages — especially to oscillator circuits and dc amplifiers. When using zeners in this application bear in mind the ratio of average to peak zener current. A figure of 0.6 is satisfactory.

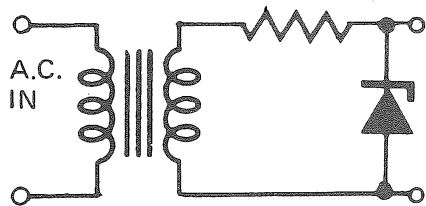


FIG. 13



FIG. 14

Zener diodes may also be placed in the primary side of a step-down (or step-up transformer). When connected in this manner the diodes will regulate all associated secondary windings. The arrangement does require high voltage rated zeners and is sometimes rather costly — however it is often used when high voltage secondary supplies need rudimentary stabilization.

Where power consumption is a prime consideration on ac power circuits, the zener load resistor (R_s) can be replaced by an inductance or capacitance. The device selected should have a reactance approximately the same as the calculated value for R_s at the supply frequency.

Oscilloscope calibrator

A single zener diode may be used as an inbuilt oscilloscope voltage calibrator that is independent of line voltage variations. Figure 16 shows how simply this facility may be incorporated in practically any oscilloscope. A selected 10 volt zener may be used to provide a calibration voltage of one volt per division.

FIG. 15

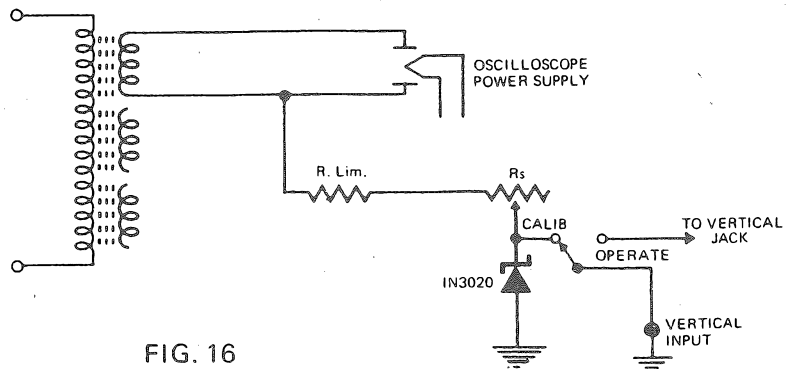
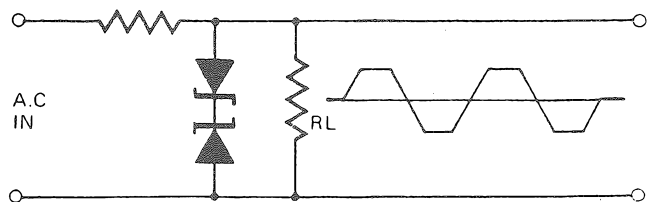


FIG. 16

Zener noise voltages

As with neon regulator tubes, zener diodes generate noise voltages. With zener diodes, these voltages are associated with junction avalanche effects, and may vary between 10 μV and 1 mV depending upon the zener type and voltage rating.

But unlike neon regulator tubes — where the incorporation of parallel filter capacitors is an excellent (if unintentionally) way of making a relaxation oscillator — a zener diode may be suppressed by adding parallel capacitance of 0.01 to 0.1 μF . This will reduce the noise voltage by a factor of at least 10 and yet maintain completely stable operation. (Figs. 17 & 18).

The zener diode as a filter

The zener diode will respond to ripple voltages in much the same manner as it does with slow voltage variations. It has a very low dynamic impedance and thus reacts in much the same way as a filter capacitor.

Excellent power supply filtering can be obtained by connecting a zener diode (having a zener voltage equivalent to the ripple trough) across the load. In most circuit applications this will be as effective as adding a smoothing capacitor of several thousand microfarads, and will provide a considerable reduction in the level of ripple superimposed on the dc output.

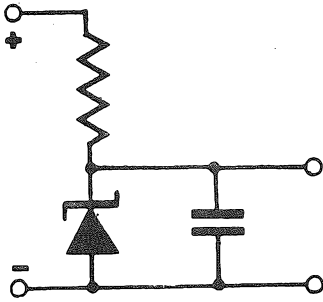


FIG. 17

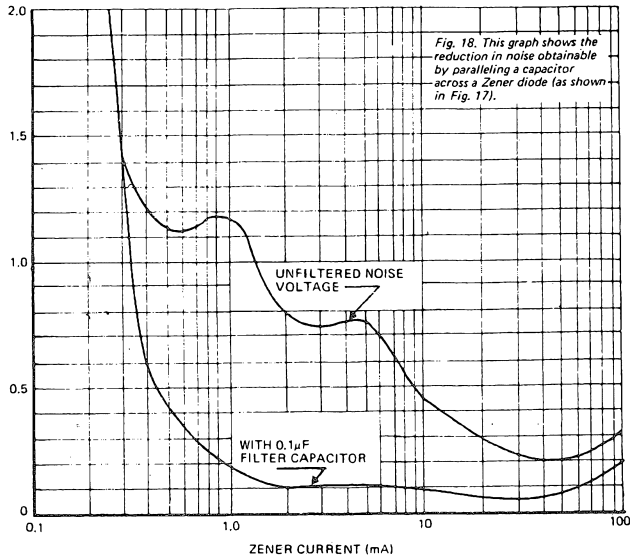


Fig. 18. This graph shows the reduction in noise obtainable by paralleling a capacitor across a Zener diode (as shown in Fig. 17).

Increasing power handling

Parallel connection is one way of increasing the power capabilities of zener diodes. But a simple parallel connection (Fig. 19) is not practicable, for the zeners will rarely be sufficiently well matched to conduct at exactly

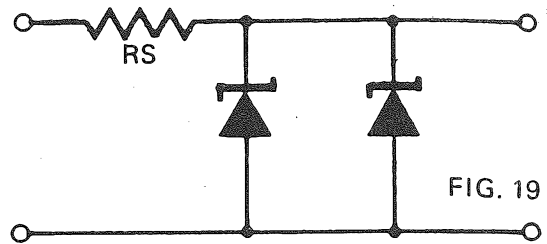


FIG. 19

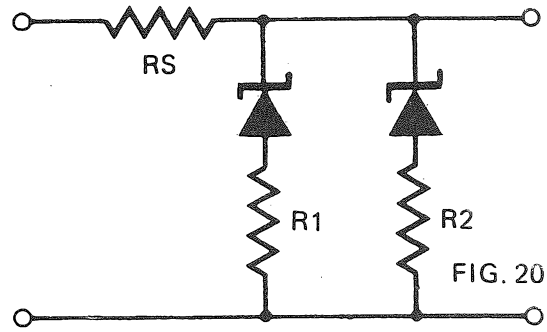


FIG. 20

the same voltage. But by including very low resistance trimming resistors, the current levels can be matched so that both zeners carry substantially equal amounts of the total current. (Fig. 20).

Although, as shown above, zener diodes can be paralleled in order to increase their load carrying capacity it is usually more practicable to use a series shunt transistor circuit with a zener diode providing a voltage reference. This configuration will not only improve the power handling capability by a factor of ten or so, it will also improve the regulation of the circuit by an amount equal to the current gain of the transistor.

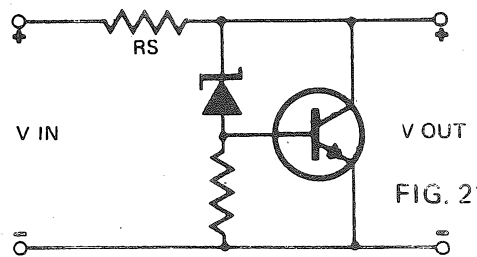


FIG. 21

A simple zener controlled shunt regulator is shown in Fig. 21.

The shunt regulator is very suitable for experimental and instructional use as it is totally short circuit proof. But, since maximum transistor current flows at zero load (and vice versa), it is also very inefficient, and for applications where power availability is limited the series transistor configuration is preferable.

A series regulator is shown in Fig. 22. In this circuit the zener diode establishes a reference voltage for the series transistor, which, in effect, operates as an emitter follower. Thus the emitter voltage is held within a few tenths of a volt of the base potential (which is determined by the zener diode).

Thus the transistor acts as a series element to absorb voltage variations. All load current flows through this series transistor. The power handling ability of this type of supply is determined entirely by the number and type of transistors used (and the ability of the heat sink to remove heat). Figure 23 shows the regulation obtained from the simple circuit of Fig. 22 which uses a 4.7V 1/4W zener and a 1k series resistor. The regulation may be improved by a factor of 10 by substituting a low dynamic resistance 4.7V 3.5W zener (we used an IR 1N1589) for the 1/4W device.

This circuit (Fig. 22) can provide a variable voltage regulated output by connecting a 1k potentiometer across the zener diode. The variable reference voltage is then applied to the base of the series reference transistor. However, this modification reduces the degree of regulation due to the shunting effect of the potentiometer. A better system is to switch different zener diodes into the circuit for different voltage outputs.

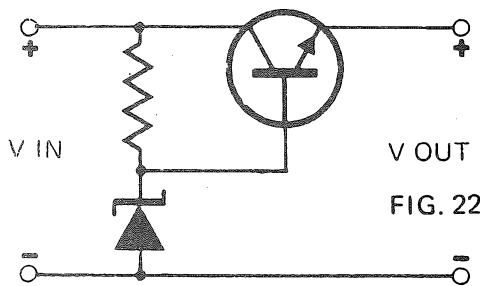


FIG. 22

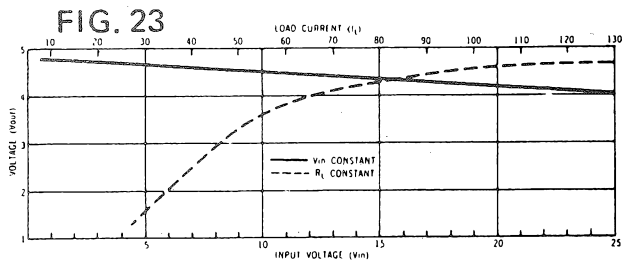


FIG. 23

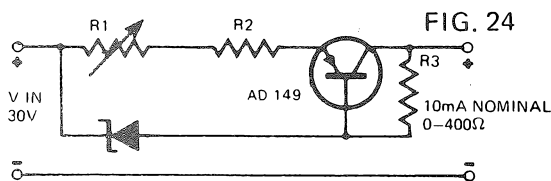


FIG. 24

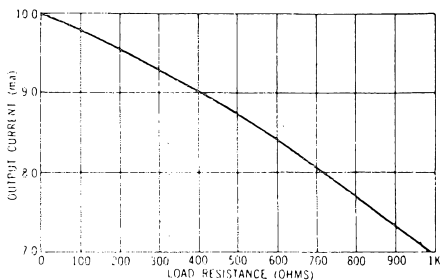


FIG. 25

Constant current regulation

A simple zener-regulated constant current supply can be built using a single transistor as a variable series resistor. Figure 24 shows how it is done. Two circuit paths exist; one through the zener diode which is in series with the bias resistor, and the other through R1, R2 and the AD149 series transistor. Any change in the current through R3 causes a change in bias, the series transistor thus changes resistance to correct the current flow. In operation the current will remain substantially constant (within about 10%) from a short circuit to a 400 ohm output load. A graph of the output characteristics is shown in Fig. 25.

Logic

In many ways a zener diode resembles a switch, and is therefore often used in computer and instrumentation logic circuitry.

The advantage of zener diodes for this purpose is their extremely rapid operation when switching around their avalanche point. Whereas germanium or silicon diodes are limited to data rates of less than 2.5MHz (due to storage of minority carriers) zener diodes switching about their avalanche point have switching times practically equal to their relaxation time. For silicon this is 10^{-9} seconds.

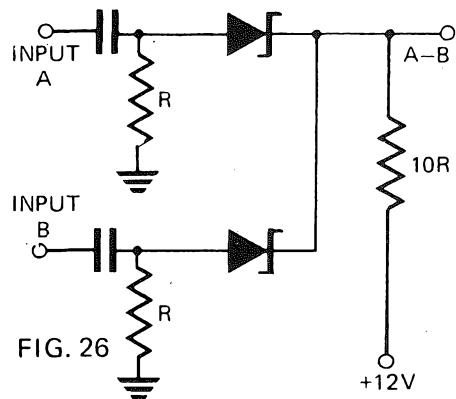


FIG. 26

A typical AND gate, using two 6.8V zener diodes is shown in Fig. 26. The positive 12 volt bias is applied to both cathodes through the common load resistor, maintaining the diodes in avalanche condition. If a positive pulse is now applied to input A, the associated zener will be gated out of the avalanche region. But due to the low impedance of the parallel connected zener (associated with input B), the output remains clamped at 6.8volts. The same circuit conditions apply if a pulse is applied solely to input B. However, if a pulse arrives simultaneously at points A and B, both zeners will be gated out of the avalanche region and the output voltage will rise from 6.8 volts to the 12 volt supply potential during the time that the pulse is coincident at the two inputs. Thus a positive pulse is produced coincident with the input pulse. Negative input pulses will not affect the circuit.

Figure 27 shows how a similar technique may be used to provide an OR function. In this configuration a pulse applied to *either* input will produce a pulse at the output.

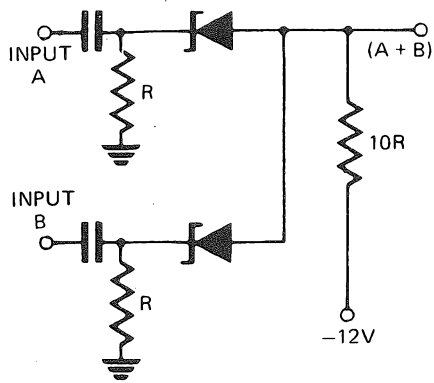


FIG. 27

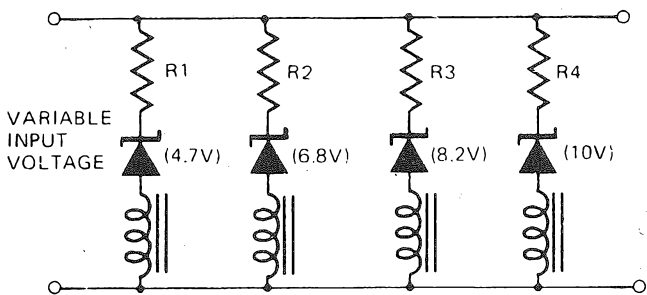


FIG. 28

Sorting

In the circuit shown in Fig. 28 the relays will be progressively energized as the input voltage increases. This circuit is often used for voltage controlled sorting. The relay contacts may be arranged to open chutes and illuminate indicator lamps for rapid sorting.

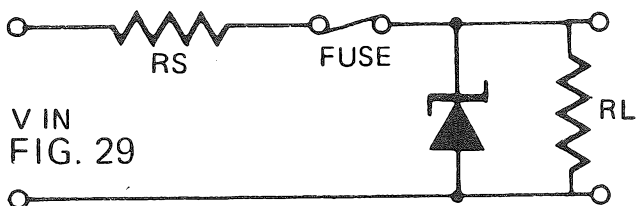


FIG. 29

Overvoltage protection

The voltage sensitive characteristic of zener diodes can be combined with the current sensitive characteristic of fuses to protect circuit components from overvoltage surges, whilst at the same time eliminating the 'nuisance' fusing that occurs when a fuse too close to the operating current is used. By connecting a zener diode of the correct voltage rating across the load, a fuse adequate to carry the normal load operating current for long periods may be used. But if the input voltage increases — and so exceeds the zener breakdown voltage — the zener diode will conduct. The sudden increase in current will blow the fuse practically instantaneously. (Fig. 29). A similar circuit may be used in conjunction with a circuit breaker rather than a fuse.

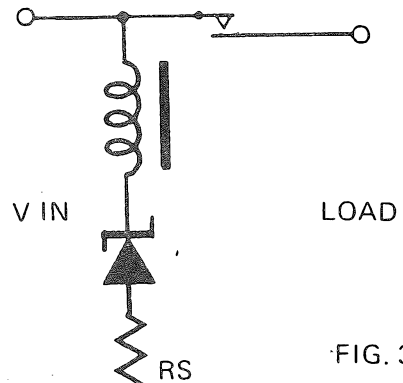
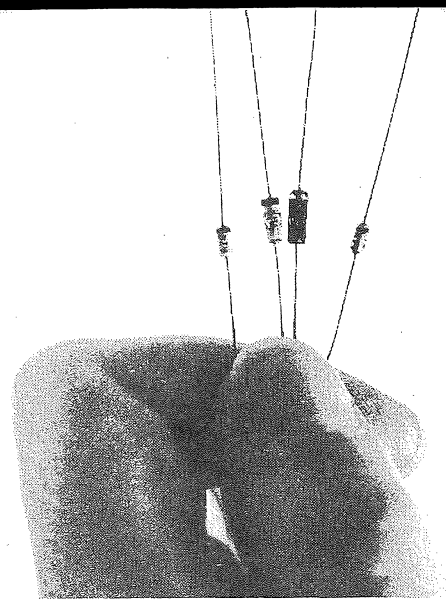


FIG. 30

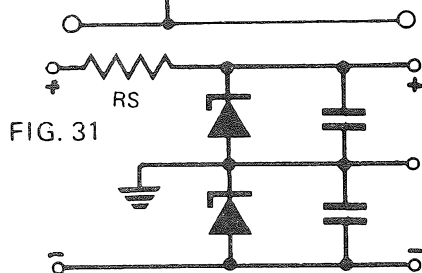


FIG. 31

Undervoltage protection

In some applications it may be necessary to disconnect a load from the mains supply if the supply voltage falls below critical level. A simple circuit that will provide this function is shown in Fig. 30. The series resistance R_1 is chosen so that at normal operating voltages the zener diode is broken down and sufficient current flows to hold the relay closed. When the supply voltage falls below the desired level, the zener ceases to conduct and the relay drops out. The addition of the zener diode to this circuit provides an accurate reference point, increasing reliability and eliminating the need for specially selected relays for different voltages.

Dual voltage supply

Most analogue circuitry needs a dual power supply — (one positive and one negative with respect to zero). The useful, but little known circuit shown in Fig. 31 can supply a dual output of balanced or unbalanced voltage from a single ended power supply. Zener diodes should be chosen to suit the voltages required.

Full details of operating characteristics of zener diodes can be obtained from most semiconductor manufacturers.

MODERN CRYSTAL OSCILLATORS

Roger Harrison looks at circuits for Radio Amateurs

Crystal oscillators in one form or another, are fundamentally associated with virtually all transmitting and receiving equipment. Basic circuitry and circuit techniques, and the fundamentals of quartz crystals are discussed at various length by both the ARRL and RSGB handbooks, Pat Hawker's 'Amateur Radio Techniques', the various VHF handbooks by Jessop (RSGB) and Tilton (ARRL) as well as 'The Radio Handbook' by Bill Orr (Editors and Engineers). A useful and more recent discussion on the subject of crystals and crystal oscillators is contained in the 'Ham Notebook' from the editors of the American journal 'Ham Radio'. For a deeper appreciation of the subject, references (1) to (4) are recommended.

Basic solid state crystal oscillator circuit techniques are by now well established, most circuits being adaptations of the well-known vacuum tube technology such as the Pierce, Hartley, Clapp and Butler oscillator and use both bipolar and FET devices. Whilst these circuits basically fulfil

their intended purpose, there are many applications which require something different or where performance needs to be reliably characterised.

Presented here are a variety of circuits, for a range of applications from LF through the VHF range, that are not commonly found in current amateur use or literature.

MODES OF OPERATION

A point not often appreciated, or just forgotten, is that quartz crystals can oscillate in a *parallel* resonant mode and a *series* resonant mode. The two frequencies are separated by a small amount, typically 2-15 kHz over the frequency range. The series resonant frequency is *lower* in frequency than the parallel. A crystal specified and calibrated for use in the parallel mode may be satisfactorily used in a series resonant circuit if a capacitor equal in value to its specified load capacitance (usually 20,30, 50 or 100 pF) is connected in series with the crystal. Sadly, you can't invert the process for series resonant crystal in parallel mode circuits. The series mode crystal will oscillate higher than its

calibrated frequency in this case and it may not be possible to capacitively load it down sufficiently.

Overtone crystals operate in the *series* mode usually on the third, fifth or seventh overtone, and the manufacturer normally calibrates the crystal at the overtone frequency. Operating a crystal in the parallel mode and multiplying the frequency three or five times produces quite a different result from operating the same crystal in the series mode on its third or fifth overtone. When ordering overtone crystals avoid confusion and specify the frequency you want, *not* the apparent fundamental frequency. Reference (4) makes this point quite clear.

Fundamental crystals in the range 500 kHz to 20 MHz are usually specified for parallel mode operation but series mode operation can be requested. For low frequency crystals, up to 1 MHz, either mode can be specified. Overtone crystals generally cover the range 15 MHz to 150 MHz.

WIDE RANGE or APERIODIC OSCILLATORS

Oscillators that do not employ tuned circuits can be very useful, whether they are simply used as 'crystal checkers' or some other purpose. Particularly for LF crystals, tuned circuits can be bulky. However, they aren't without their traps. Some crystals are prone to oscillation on unwanted modes, particularly the DT and CT cut crystals used for LF quartz oscillators. It is wise to check that the output is on the correct frequency and no mode instability is evident. Reducing feedback at the higher frequencies usually cures this. In extreme cases, the idea has to be abandoned and an oscillator having a tuned circuit used instead, (LF crystal oscillators are discussed later).

The first circuit is an emitter-coupled oscillator, a version of the Butler circuit. The basic circuit first appeared in VHF Communications in 1970 (p.240) as portion of a VHF-UHF calibration spectrum generator.

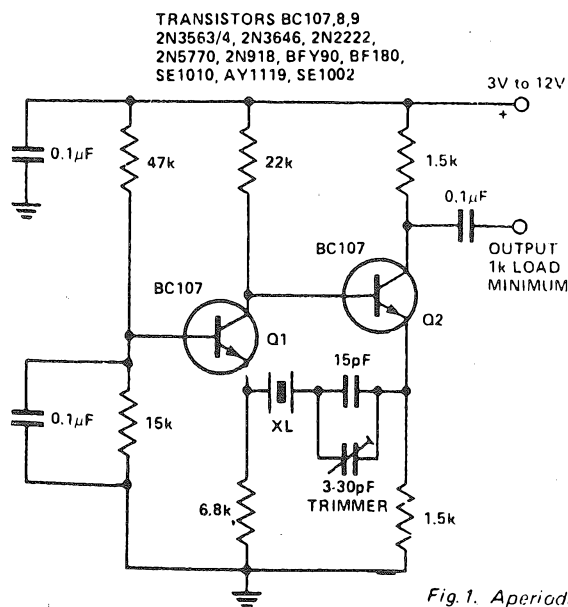


Fig. 1. Aperiodic Butler oscillator (series mode)

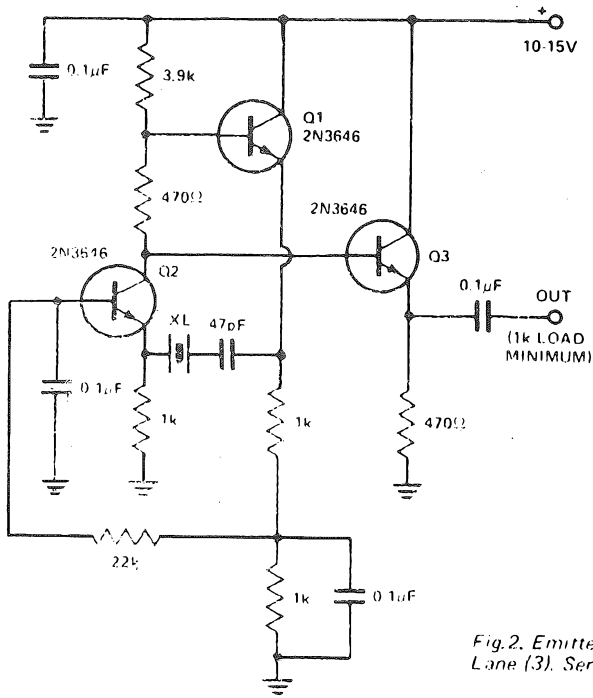


Fig. 2. Emitter-coupled oscillator — after Lane (3). Series mode.

Versions have subsequently been published in the 'VK5 Bulletin' (S.A. Div. WIA) in 1972 and 6UP, August issue, 1974. Lane (3) discusses a variation of this circuit (Fig. 2).

The output of the circuit in Fig. 1 is essentially sine wave; reducing the emitter resistor of Q2 increases the harmonic output. By doing this, a 100 kHz crystal produces good harmonics through 30 MHz. It is a series mode circuit.

A variety of transistors may be used. For crystals above 3 MHz, transistors with a high gain-bandwidth product are recommended. For crystals in the 50 kHz to 500 kHz range, transistors with high LF gain, such as the 2N3565 are recommended. Also, for crystals in this range, permissible dissipation is usually less than 100 microwatts and amplitude limiting may be necessary. Low supply voltage, consistent with reliable starting, is recommended. Modifying the circuit by the addition

of diodes — as shown in Fig. 3 — is a better method, and starting performance is improved. The circuit will oscillate up to at least 10 MHz with appropriate transistors and emitter resistor values. An emitter follower or source follower buffer is recommended. Similar comments to the above apply to Fig. 2. An emitter follower buffer is included in this circuit. Both circuits are slightly frequency sensitive to power supply voltage changes and load variations. A load of 1 k or greater is recommended.

TTL ICs can be used in crystal oscillator circuits but many published circuits have poor starting performance or suffer from non-repeatability owing to wide parameter spreads in ICs. The circuit in Fig. 4, is by K1PLP from QST, Feb. 1974 (5) and is after Weggeman (6). This circuit has been tried by the writer over the range 1 MHz to 18 MHz and can be recommended. It is a series

mode oscillator and suits AT-cut crystals. The output is about 3 volts peak to peak, square wave up to about 5 MHz beyond which it is becomes more like half-sine pulses. Starting performance is excellent, often a critical factor with TTL oscillators.

LOW FREQUENCY CRYSTAL OSCILLATORS

Crystals in the range 50 kHz to 500 kHz require special considerations not encountered with the more common AT or BT cut HF crystals. The equivalent series resistance (which determines 'activity' — that figure of merit of days of old) is much greater and their permissible dissipation is limited to less than 100 microwatts, preferably 50 microwatts or less.

The circuit in Fig. 5, is a series mode oscillator described by Lane (3). It has the advantage of not requiring a tuned circuit, and has a choice of sine or square wave output. For crystals in the range 50-150 kHz, 2N3565 transistors are recommended although the author has found BC107's satisfactory. Either type will suffice for crystals in the range 150 kHz to 500 kHz. If you find the crystal will not start reliably, most likely the crystal has a very high equivalent series resistance, in which case increase R1 to 270 ohms and R2 to 3.3 k (as recommended by Lane). For square wave operation, C1 is 1 μF (or a value close to, or above it). For sine wave output, C1 is not in circuit. Amplitude limiting is unnecessary. Sine wave output is about 1 V rms, square wave output about 4 V peak to peak.

The circuit in Fig. 6 is also described by Lane (3) and can be recognised as a modified form of the Colpitts oscillator, with the addition of resistor Rf to control feedback (it works the same way as Eno's). Capacitors C1 and C2 should be reduced by preferred values as the frequency is increased. At 500 kHz, values for C1 and C2 should be around 100 pF and 1500 pF

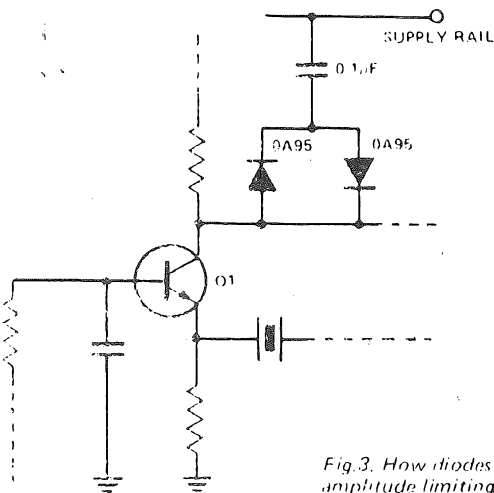


Fig. 3. How diodes are used for amplitude limiting.

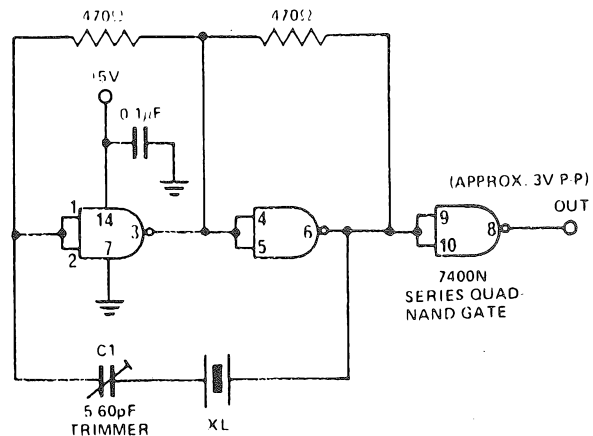


Fig. 4. Reliable TTL crystal oscillator.

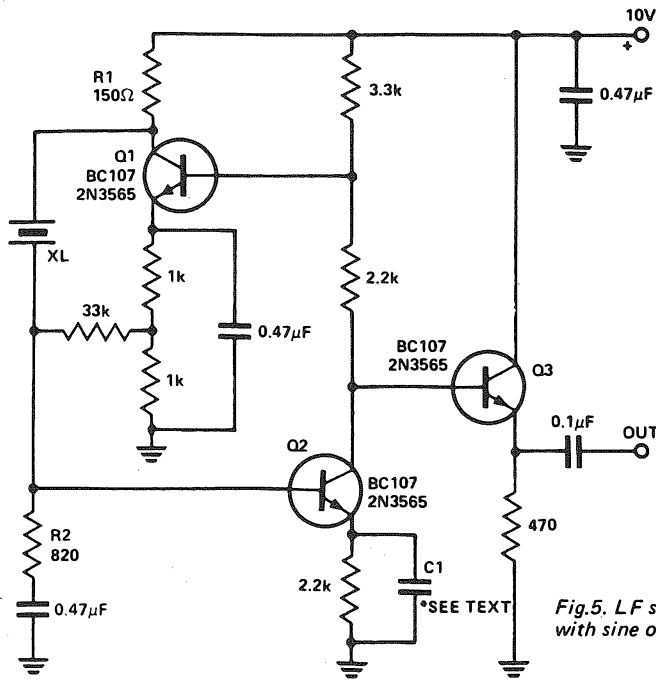


Fig.5. LF series mode oscillator with sine or squarewave output.

respectively.

The circuit as shown gives sine wave output with the second harmonic about 40 dB down (or greater). This can be reduced by careful trimming of R_f and C_1 . Note that, at the reduced level of feedback necessary to achieve this, it takes some 20 seconds for the oscillator to reach full output. Output is about 2 to 3 volts peak to peak.

If you need an output rich in harmonics, the simple addition of a 0.1 μF capacitor across the emitter resistor will achieve this. Output then rises to about 5 V peak to peak. Power supply voltage can be reduced in this case to lower crystal dissipation.

Other transistors can be used, but bias and feedback may have to be adjusted. For cantankerous crystals

determined to oscillate in modes other than those you wish, the circuit of Fig.7 is recommended. Feedback is controlled by tapping down the collector load of Q1. Amplitude limiting is necessary to keep the crystal dissipation within limits. For 50 kHz crystals the coil should be 2 mH and its resonating capacitor 0.01 μF . Output is about 0.5 V_{RMS} , essentially sine wave. The use of an emitter follower or source follower buffer is recommended. If a parallel mode crystal is used the 1000 pF capacitor shown in series with the crystal should be changed to the crystal's specified load capacitance (usually 30, 50 to 100 pF for these crystals).

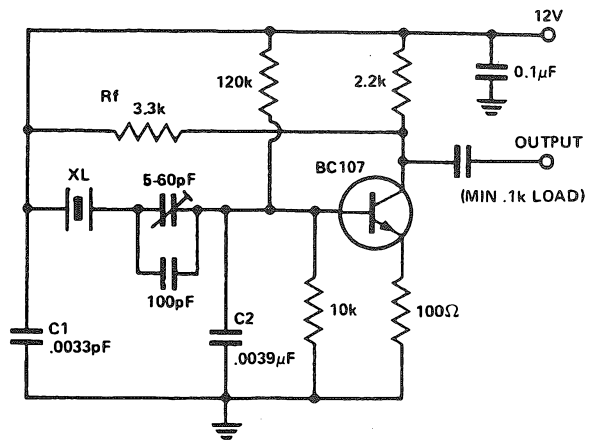


Fig.6. Parallel mode LF oscillator.

HF CRYSTAL OSCILLATOR CIRCUITS

Solid state circuits for the popular AT-cut HF crystals are legion. However, results aren't always what one would expect. Most fundamental crystals up to 20 MHz are usually specified for parallel mode operation. However, such crystals can be used in series mode oscillators by putting the specified load capacitance in series with the crystal as mentioned previously. Both types of circuit are detailed here.

A useful oscillator for the range 3 to 10 MHz that does not require a tuned circuit is given in Fig. 8 (a). It is, of course, the same circuit as Fig.6. The circuit can be used down to 1 MHz if C_1 and C_2 are increased to 470 pF and 820 pF respectively. It can be used up to 15 MHz if C_1 and C_2 are reduced to 120 pF and 330 pF respectively. This circuit is recommended for non-critical applications where high harmonic output is wanted, or not a consideration.

The addition of a tuned circuit as in 8(b) reduces harmonic output considerably. A tuned circuit with as high a Q as possible is recommended. In a 6 MHz oscillator, I have obtained the following results. With a coil Q of 50 the 2nd harmonic was 35 dB down. With a Q of 160, it was -50 dB! Resistor R_f can be adjusted (increase slightly) to improve this. The output is also increased with a high Q coil. As previously noted, with reduced feedback it takes some tens of seconds to reach full output from switch on, however, frequency stability is excellent.

Operation at other frequencies is accomplished by changing the capacitors and coil appropriately.

This circuit (Fig. 8) can also be turned into a very effective variable crystal oscillator. A small inductance is placed in series

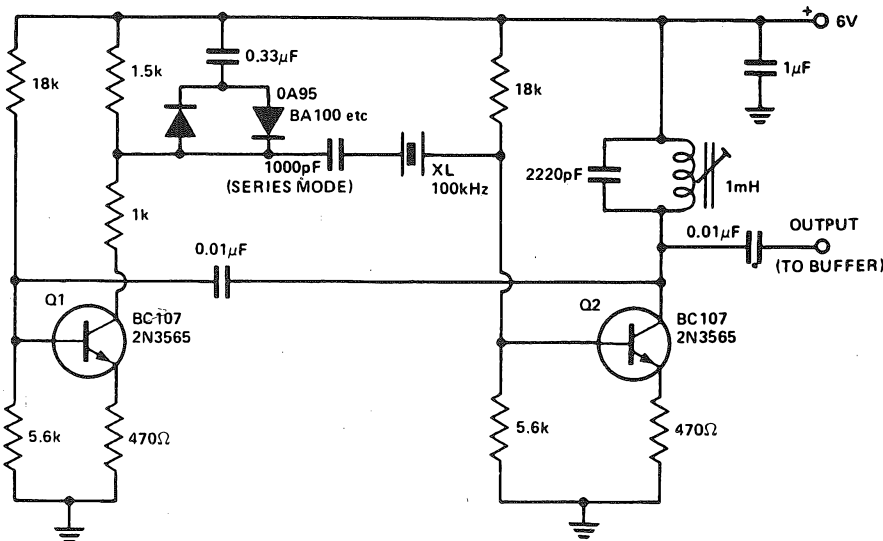


Fig.7. 100 kHz crystal oscillator (with tuned circuit).

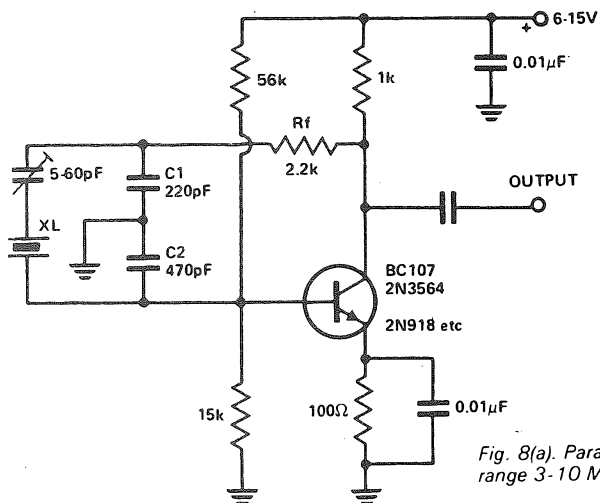


Fig. 8(a). Parallel oscillator for the range 3-10 MHz.

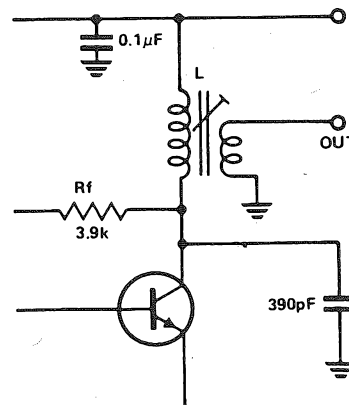


Fig. 8(b). Adding a coil to the circuit shown in Fig. 8(a).

with the crystal and one of the capacitors in the feedback circuit is made variable. An ordinary two-gang 10-415 pF (or thereabouts) broadcast tuning capacitor will do the job nicely. Both gangs are paralleled. The tuning range depends on the crystal used, the inductance of L1 and the frequency. A greater range is usually obtainable with the higher frequency crystals. Stability is excellent, approaching that of the crystal.

Another variation of this circuit is shown in Fig. 10. This circuit may allow more 'pull' on the crystal, but stability is poorer. For both Fig's 9 and 10 the trimmer is to set the nominal frequency at some position of the tuning capacitor. For both circuits also, especially for Fig.10, the output varies across the tuning range.

A VHF OSCILLATOR-MULTIPLIER

The circuit in Fig.11 is a modification of the 'Impedance Inverting' overtone oscillator discussed by Rankin (4), who also describes a similar circuit. Normally, with the impedance inverting circuit, the collector is either untuned or grounded for RF. The collector can be tuned to twice or three times the crystal frequency. To reduce the output at the crystal frequency, a double tuned circuit is recommended. DO NOT tune the collector to the crystal frequency, otherwise the circuit will oscillate at a frequency not controlled by the crystal. It is advisable to keep the collector lead as short and direct as possible.

Results with this circuit are excellent. All outputs other than the wanted output were at -60 dB or greater. Noise output is at least 70 dB below the wanted output. It makes an excellent conversion oscillator for VHF/UHF converters. Almost 2 V of RF is available at the hot end of L3

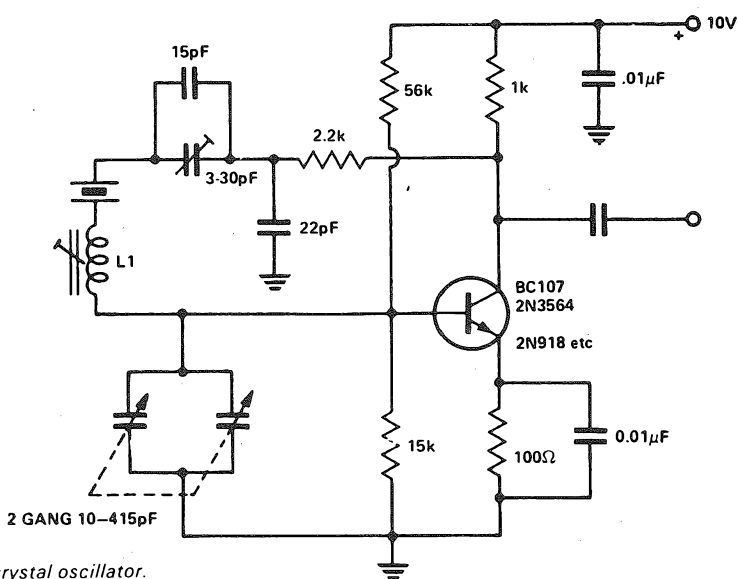


Fig. 9. Variable crystal oscillator.

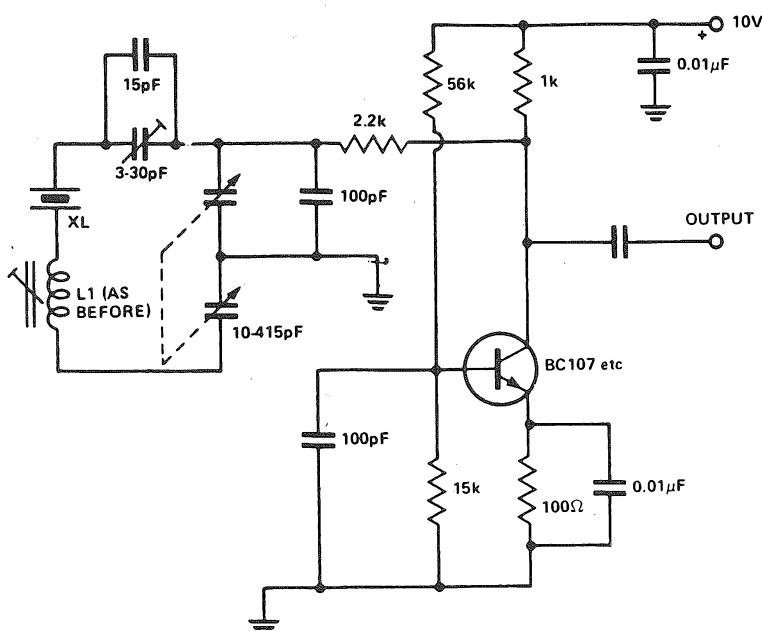


Fig. 10. Alternative version of circuit shown in Fig.9.

(author's prototype at 30 MHz). A Zener regulated supply is recommended. As indicated on the diagram, different circuit values are necessary for different transistors. Strays in individual construction may also necessitate variations. L1 can be used to pull the crystal onto frequency.

Slight variations in frequency (about 1 ppm) occur when tuning L2 and L3 and also with load variations. However, in practise, these turn out to be of no consequence.

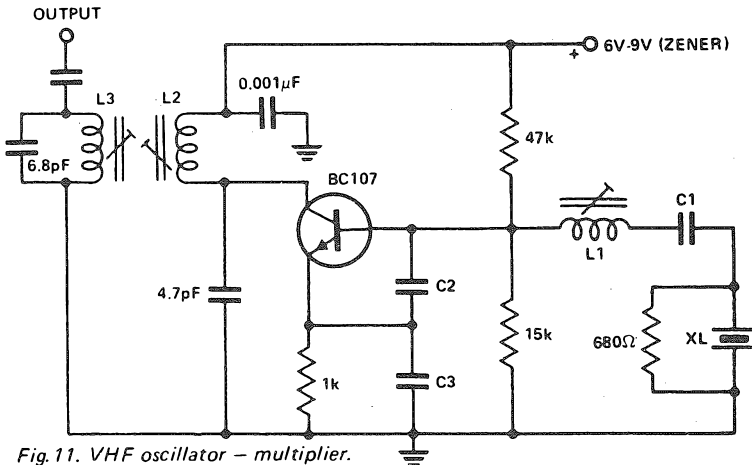


Fig. 11. VHF oscillator - multiplier.

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- (1) 'Radio Transmitters', L. Gray & R. Graham (McGraw-Hill)
- (2) 'Electronic Fundamentals & Applications', J. D. Ryder (Pitman)
- (3) 'Transistor Crystal Oscillators to Cover Frequency Range from 1 kHz to 100 MHz' by M. Lane, Australian Post Office Research Laboratories, Report No. 6513.
- (4) 'Overtone Operation of Quartz Crystals' D. Rankin (VK3QV), Amateur Radio, March and May 1967.
- (5) 'A TTL Crystal Oscillator', K1PLP, QST February 1974, p.34.
- (6) 'IC-Compatible Crystal Oscillator', The Electronic Engineer, May 1969.

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L2/3 = NEOSID, DOUBLE ASSEMBLY 7300 CAN, TWO 722/1 FORMERS, F29 SLUGS, WOUND WITH 5 TURNS, 0.63 mm ENAMEL, CLOSEWOUND

C1 = 33 OR 39 pF

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L1 = 20 TURNS 0.4 mm AS ABOVE
L2/3 = AS ABOVE
C1 = 56pF

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L1 = 24 TURNS 0.4 mm AS ABOVE
L2/3 = 9 TURNS 0.63 mm AS ABOVE
C1 = 68pF OR 100pF

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43mHz	15pF	10pF
38mHz	22pF	18pF
65mHz	18pF	12pF
43mHz	33pF	18pF
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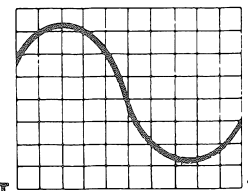
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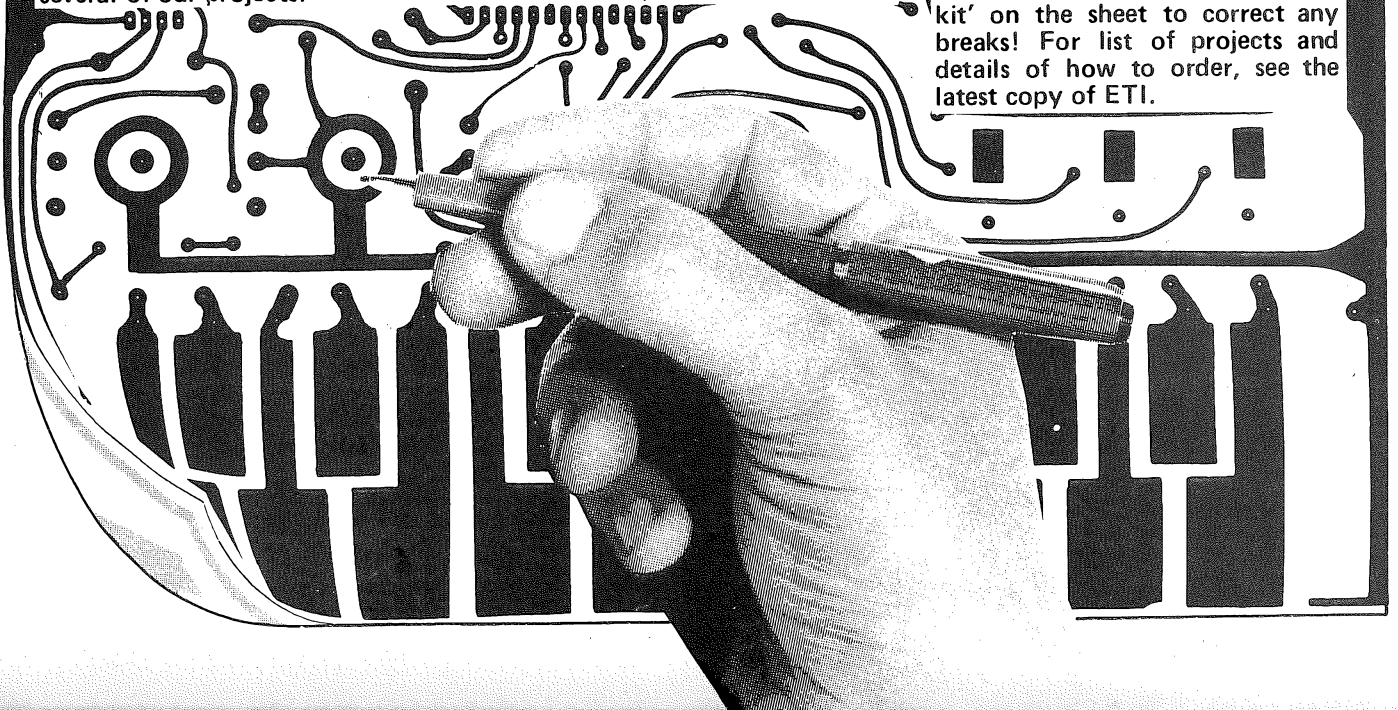
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HIGH VOLTAGE FOR LOW COST

Simple diode-capacitor networks can be cascaded to deliver any desired voltage.

If you need a power supply for a high-voltage low-current application, your best bet is probably the voltage multiplier circuit. It's inexpensive. It's simple. And you can get any voltage you want by cascading multiplier stages. The voltage is limited only by the ratings of the components you use.

A voltage-multiplier circuit contains diodes and capacitors, with the devices connected to develop a dc output that is a multiple of the peak or peak-to-peak input voltage. There are two major variations of the circuit: multipliers that use an even number of diodes and those that use an odd number of diodes.

The basic rectifier circuits in Fig. 11. (equations assume perfect diodes and capacitors, loads are considered light) can be combined to form a complete family of half-wave multipliers. A

full-wave multiplier can be made by combining two half-wave multiplier sections, one positive and one negative (Fig. 2). The major disadvantage of a full-wave multiplier is that the secondary side of the transformer nearest the core requires heavy insulation to withstand one-half the output voltage. Therefore inductive coupling is worse and efficiency lower than for a transformer used with the equivalent half-wave type. Thus half-wave multipliers are better for most high-voltage power supplies.

Figure 3 shows the two variations of half-wave multipliers. Each of these circuits consists of identical sections cascaded, except for the first stage in Fig. 3a. The first section of a multiplier with an odd number of diodes is a simple half-wave rectifier. The first section of a multiplier with

an even number of diodes is a half-wave doubler. A basic rule of thumb for multiplier designs is: For waveforms that are symmetrical about zero, use an even number of diodes; for asymmetrical waveforms, use an odd number.

CALCULATING THE OUTPUT VOLTAGE

The regulation of a multiplier with a load is a function of the input's source impedance, the values of the capacitors in the multiplier, the forward drop of the diodes and the turn-on and turn-off times of the diodes. The output voltage of a multiplier is approximately

$$V_{out} = N \frac{(V_1 + V_2)}{2} - \frac{N^3}{12Cf} \cdot I_{out}$$

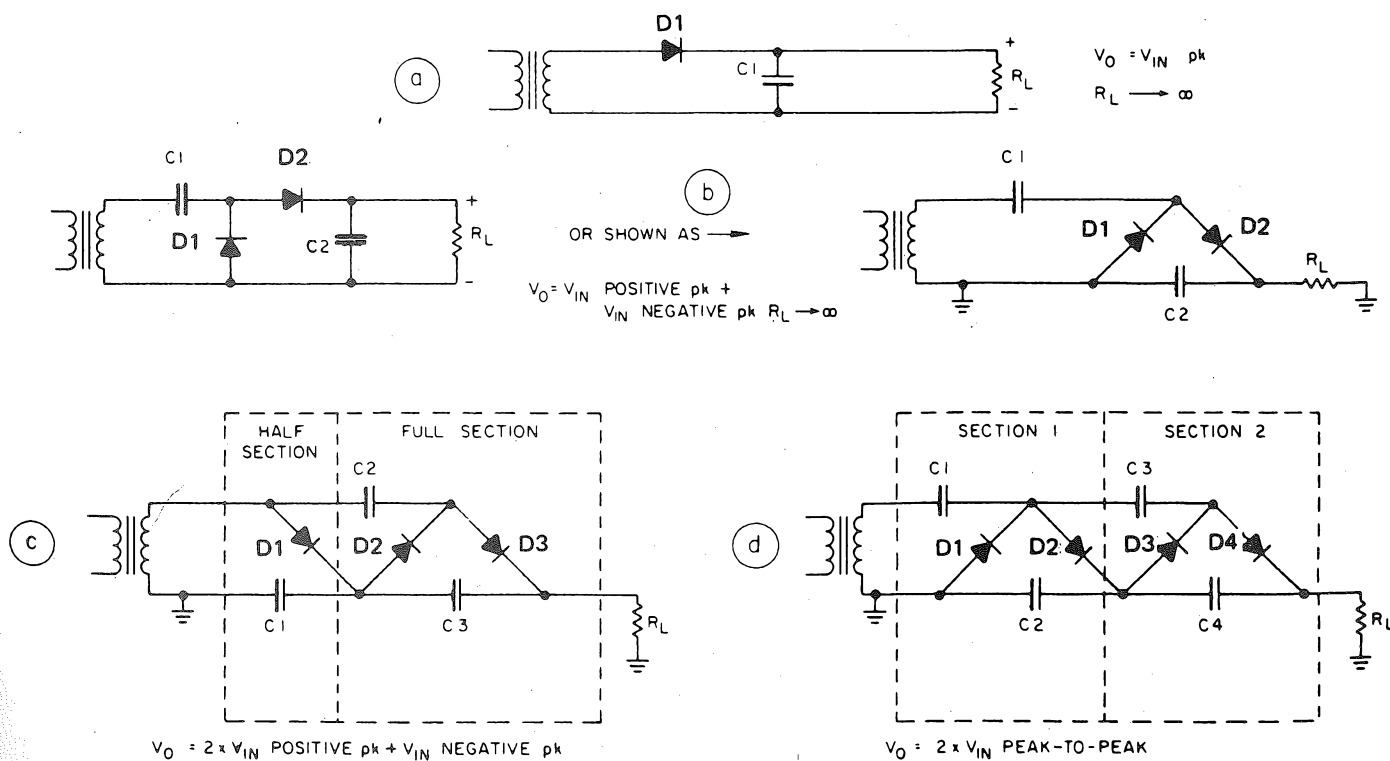


Fig. 1. The basic half-wave rectifier circuit (a), can be modified to become a voltage doubler (b). Various higher output voltages can be obtained by adding further sections (c and d).

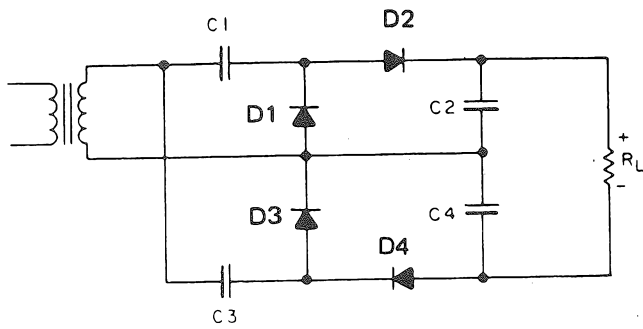
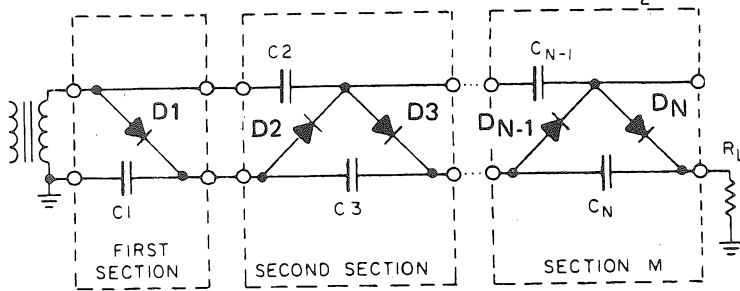


Fig. 2. The full-wave voltage quadrupler circuit requires a transformer with heavy secondary insulation.

$$M = \text{SECTIONS}$$

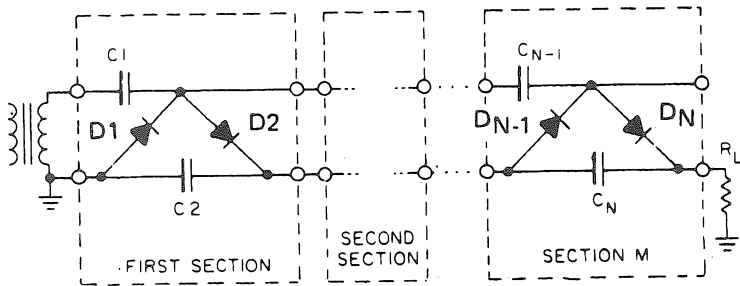
$$M = \frac{N+1}{2}$$



(a)

$$M = \text{SECTIONS}$$

$$M = \frac{N}{2}$$



(b)

Fig. 3 (a). A multiplier with an odd number of diodes works best for asymmetrical waveforms. (b) An even number of diodes is best for symmetrical waveforms.

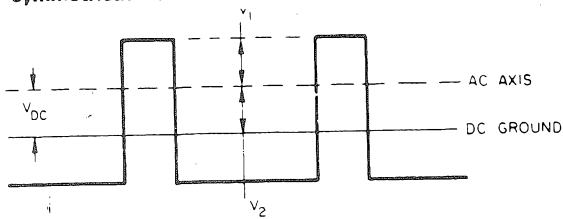


Fig. 4. A recurring waveform with a positive peak V_1 and negative peak V_2 is used as an input for the voltage multiplier circuit described in Fig. 5.

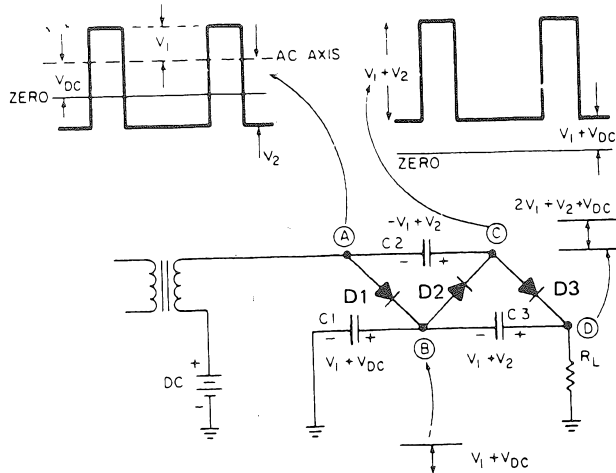


Fig. 5. The voltage waveforms at different points within the multiplier circuit with an odd number of diodes show the transformation of the pulse waveform described in Fig. 4 into a much higher dc voltage.

Here N is the number of diodes or capacitors used for circuits like those shown in Fig. 3; V_1 is the positive peak input voltage; V_2 is the negative peak input voltage; C is the capacitance in farads; f is the frequency of the input, and I_{out} is the current in amperes. This equation assumes a sufficiently large load capacitance, equal value capacitors, and ideal diodes. It will produce sufficiently accurate results for all practical purposes.

WATCH DIODE SWITCHING CHARACTERISTICS

The turn-on and turn-off times of the diodes are important if high frequencies are involved. Both turn-on and turn-off must be kept fast, its regulation and efficiency are to be maintained. The forward drop of the diodes is usually not a significant factor.

The output regulation of voltage multipliers ranges from 100 V to 5 kV per milliamp of current. Some applications use regulation schemes to control power-supply output. Some common methods are shunt dc load, rectified pulse feedback and a saturable reactor in series with the high-voltage transformer. In other applications, it is desirable to have the output voltage sag with load - with very poor regulation built into the multiplier through selection of the capacitor's value.

The output voltage of a multiplier will always have some ripple in the output. Ripple is a function of load capacitance, input frequency, multiplier impedance and input-to-output coupling.

The load capacitance acts as a filter, and the effective series impedance of the multiplier limits voltage ripple. If regulation is not a consideration or if load current is almost constant, a series resistor can be added to the multiplier output. The series resistor will act with the load capacitance as an RC filter.

The high-frequency components of the input voltage are the most easily coupled into the output. But the higher frequencies are also easier to filter at the multiplier output when necessary. The most unpredictable ripple component, though, is generated by stray capacitive coupling of the input to the output terminal. This coupling is difficult to control. The mechanical layout of the multiplier can reduce it, and if more ripple reduction is required, an electrostatic shield can be used to isolate the output area further from the input.

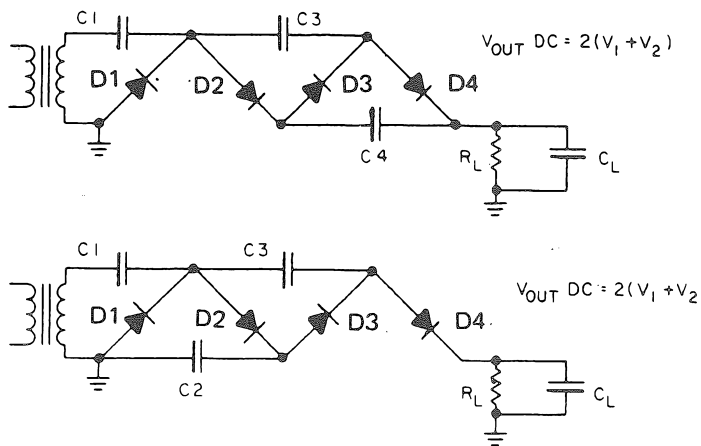
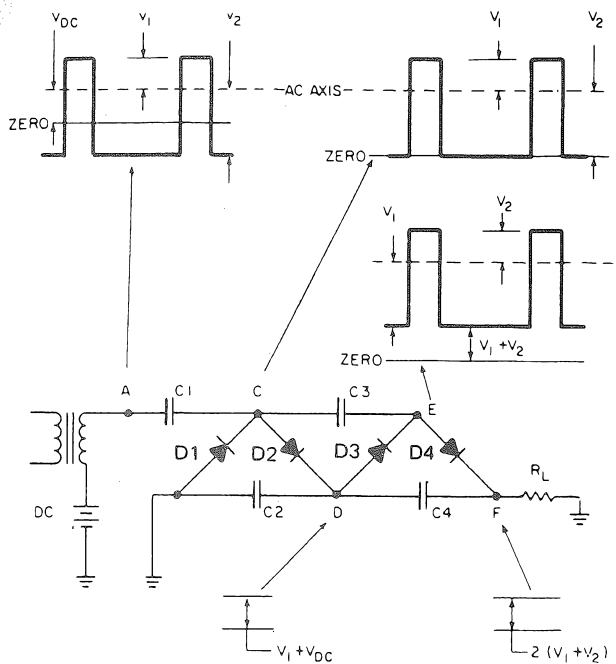


Fig. 7. To reduce component cost and count if the load is capacitive remove one of the doubling capacitors.

Fig. 6. The multiplier circuit with an even number of diodes and the same input as described in Fig. 4 produces an even larger dc output voltage than the circuit of Fig. 5.

HOW MULTIPLIERS WORK

The multiplier circuit can handle any waveform, but the three most common for multiplication are sine, pulse (or square) and trapezoidal wave. The only waveform restrictions are that the rise and fall times of the input signal be slower than the diode switching time.

In the signal in Fig. 4, V_{in} is a recurring waveform composed of the positive peak V_1 , the negative peak V_2 and an ac axis that can be displaced from dc zero by voltage V_{dc} .

Figure 5 shows the voltages at each point of a 1.5-section multiplier. The half-wave, 1.5-section multiplier (three diode) operates as follows: During the positive peak of V_{in} , diode D_1 conducts to charge C_1 to a voltage equal to $V_1 + V_{dc}$. Capacitor C_2 acts as a coupling capacitor to couple V_{in} to point C. Diode D_2 conducts on the negative voltage peak at point C when the voltage tries to become more negative than the anode of D_2 (the anode voltage of D_2 is $V_1 + V_{dc}$). Diode D_3 conducts on the positive peak at point C and charges D_3 to $V_1 + V_2$. The output, V_{out} , is the sum of the voltages on C_1 and C_3 :

$$V_{out} = V_1 + V_{dc} + V_1 + V_2 = 2V_1 + V_2 + V_{dc}$$

Only dc voltages are applied to C_1 and C_3 ; these capacitors are therefore dubbed "dc capacitors." An ac voltage is applied to C_2 , which is called an "ac capacitor." If the input voltage is symmetrical about the zero axis, the multiplier output will be three times (either) peak voltage, $V_{out} = 3V_1$. This circuit is called a tripler. If, however, the waveform is such that V_1 is much greater than V_2 , the output voltage is approximately twice V_1 . The circuit could be called a doubler. For clarity, we can use the diode count to define multiplier capability.

The operation of the four-diode multiplier — a two-section, half-wave unit — is similar to that of the three-diode multiplier (Fig. 6). Capacitor C_1 blocks the dc bias from the remainder of the multiplier and acts as a coupling capacitor to couple V_{in} to point C. Diode D_1 conducts when the negative voltage at point C becomes more negative than the anode of D_1 (the anode of D_1 is at 0 V). This causes C_1 to charge to a voltage equal to $V_2 - V_{dc}$ and simultaneously causes the positive peak at point C to reach $V_1 + V_2$.

The positive voltage at point C turns on D_2 and charges C_2 to $V_1 + V_2$. Capacitor C_3 acts as a coupling capacitor to couple the input waveform at point C to point E. Diode D_3 conducts when the cathode voltage becomes more negative than the anode voltage (the voltage at point D). The positive peak will be at a voltage equal to the charge on C_3 plus the peak voltage at point C. This positive voltage will cause D_4 to conduct and charge capacitor C_4 to $V_1 + V_2$. The output, V_{out} is the sum of the voltage on C_2 and C_4 :

$$V_{out} = (V_1 + V_2) + (V_1 + V_2) = 2V_1 + 2V_2$$

Both C_2 and C_4 are dc capacitors. Points D and F are "dc points", and C_1 and C_3 are ac capacitors. In both the odd-diode and even-diode circuits, the diode peak-inverse voltage (PIV) ratings should be at least $V_1 + V_2$. In the even-diode multiplier, C_1 should have a voltage rating of at least V_2 . In the odd-diode multiplier, C_1 should have a voltage rating of at least $V_1 + V_{dc}$. All the other capacitors should have a voltage rating of at least $V_1 + V_2$. Negative output voltages can be obtained if the diode polarities are reversed.

VARIATIONS FOR SPECIAL APPLICATIONS

For applications with a very high load capacitance, any one of the dc capacitors can be omitted in the multiplier and it will still function (Fig. 7).

While this appears to be a good way to reduce component costs and package size, consider what happens when the output terminal is arced to ground: the distribution of voltages on the diodes becomes unequal, which causes more stress on some diodes than others. The uneven distribution can cause a diode's peak inverse rating to be exceeded and a malfunction to occur. For better transient protection, leave all the capacitors in the circuit.

Many applications require a second voltage that is proportional to the output voltage. A tap at any dc point of the multiplier can be used. The ratio of the voltages can be determined if you examine the circuit up to the tap as a complete unit and the total multiplier as another.

Consider carefully the maximum average current. The multiplier current ratings are intended to keep the components cool enough to perform reliably. It will help, of course, if the high-voltage drive source has some

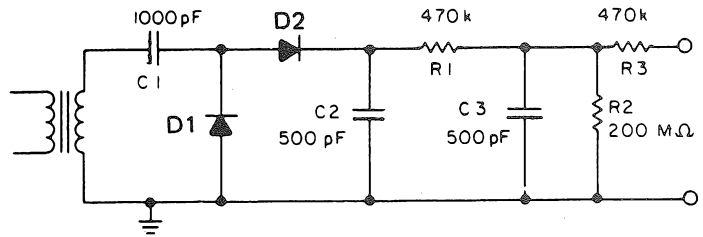


Fig. 8. A voltage doubler can be combined with a filter to provide very low ripple outputs. (The example shown gives a 30 kHz, 10 kV, 50 μ A output).

maximum-load protection that reduces the input voltage if too much current is demanded.

The multiplier must withstand all arcing, including that between the output lead and ground, and also direct shorts of the output lead to ground. The multiplier must sustain the peak current drawn by the arc or short as the internal capacitors discharge.

A resistor in series with the output lead serves two functions: (1) It reduces the Q of the oscillator circuit that is established during arcing, thus reducing considerably the stress on the diodes, and (2), it limits the peak current to a value that the diodes can handle safely. The value of this

resistance must be high enough to do the limiting job but not so high as to promote arcing around or through the resistor body or overheating at maximum current drain when the output arcs to ground.

CONSIDER THE MECHANICAL LAYOUT

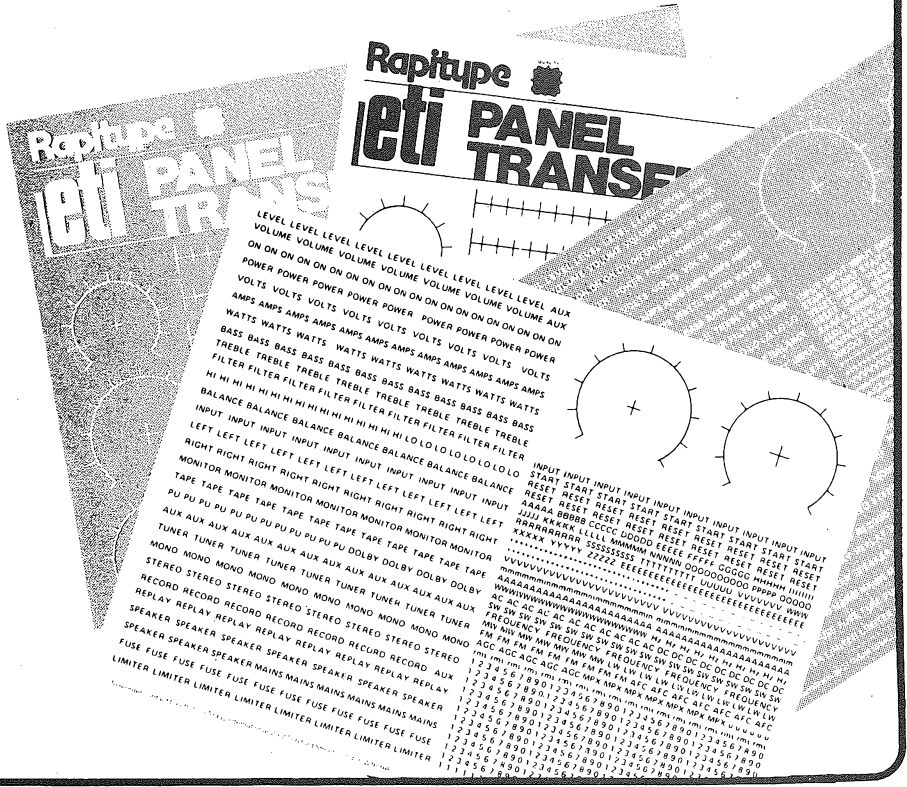
The mechanical design, mounting method and location of the multiplier can all affect current capability.

Remember that very high voltages may be involved so pay particular attention to component layout and insulation — also ensure that there are no sharp edges that might otherwise initiate corona discharge.

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ODD ODES

A diode, the electronic one way street, is a versatile component. This tiny piece of crystal engineering can rectify AC signals, limit voltage, emit light or tune your radio. Ian Sinclair explains

If you compare a resistor to a crowded road and a capacitor to a multistorey car park, then a diode is the nearest thing electronically to a one-way street. A diode has two terminals (the di-part of the name simply means two) and the current flows only when one of them, the anode, is more positive than the other, the cathode. This direction of current flow, anode to cathode, is called the forward direction and doesn't obey Ohm's Law. That means that we can't calculate how much current will flow simply by measuring the forward voltage and knowing a single figure of resistance of the diode, R . There are two features of the way in which a diode conducts which makes it quite different from a resistor. One is that current doesn't start to flow whenever the anode is positive to the cathode, only when the voltage is greater than about 0.5 V (for silicon diodes) or 0.15 V (for germanium diodes). The other feature is that, once the diode is conducting, its resistance drops as the current increases. The drop in resistance is so great that the voltage across a forward conducting diode is almost constant, around 0.55 V, even if the current changes considerably. For silicon diodes, a very useful rule of thumb is that the voltage changes by only 60 mV for a tenfold change of current. This means, for example, that if the voltage across a diode is 0.55 V when 1 mA is flowing, then increasing the diode current to 10 mA will raise the voltage by only 60 mV to 0.61 V. If the diode obeyed Ohm's Law, then a tenfold increase in current would cause a tenfold increase in voltage. In our example, a resistor which had a voltage of 0.55 V across it with 1 mA flowing (a 550R resistor) would have 5.5 V across it when 10 mA flowed. Diodes just don't behave that way.

Characteristics

If we can't use Ohm's Law then, what do we do? The answer is that we have to use characteristics, graphs which show how much current flows at each value of voltage. A full set of characteristics for a diode is quite an impressive sheaf of documents, but the two that are of most interest to us are the forward characteristic and the reverse characteristic. The forward characteristic shows how much current will flow at each value of forward voltage and at what voltage current can be expected to start flowing. The reverse characteristic shows how

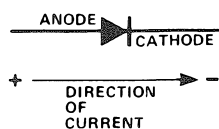


Fig. 1. Symbol for a diode. The arrowhead on the symbol shows the conventional direction of current (+ to -) through the diode.

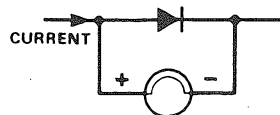


Fig. 2. Measuring the forward voltage for a conducting diode. This is always around 0V5 for a silicon diode, 0V2 for a germanium diode.

much reverse or leakage current will flow when the diode is reverse biased (cathode positive, anode negative) to various voltages. This reverse characteristic usually has a turnover (Fig. 3) and in the normal use of a diode we try to avoid applying a reverse voltage large enough to reach this turnover point. Why? Well, unless there's enough resistance in the circuit to make sure that the current which can flow in the reverse direction is very small, enough power will be dissipated to overheat the diode and destroy it. The power converted to heat (in milliwatts) is given by volts \times milliamps. If the diode can just safely pass 20 mA in the forward direction, when the forward voltage, is, say, 0V6 then the power it can

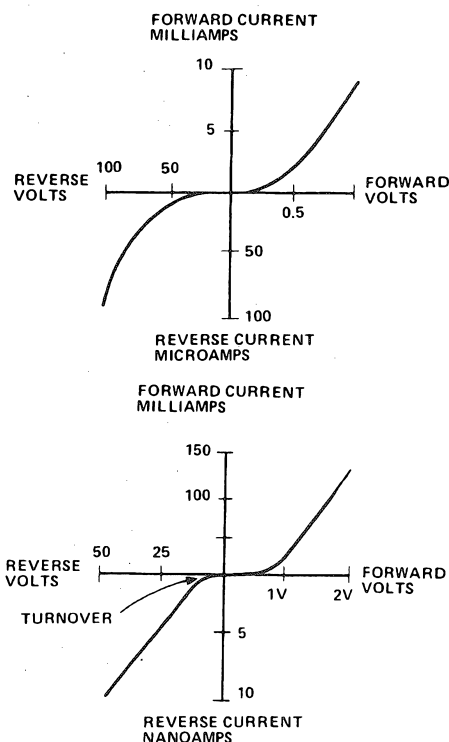


Fig. 3. Forward and reverse characteristics plotted on one graph. (a) Germanium diode, (b) Silicon diode. Notice that the scales for reverse voltage and current are not the same as the scales for forward voltage and current. This has to be done so as to get the two different characteristics on the same graph.

handle is $0.6 \times 20 = 12$ mW. In the reverse direction, if the turnover is at -20 V, then the power which has to be dissipated at 20 mA is $20 \times 20 = 400$ mW — and it won't like it!

Why Diodes Do It

That's what a diode does, but why does it do it? The answer to that question is not so easy, because it needs some understanding of how materials are formed from atoms and molecules. Let's try to get by with a simple explanation on the understanding that there's a lot more to it. First of all, the materials that are used for making diodes or transistors are solid crystals. Crystals of a given material always have the same angles between faces, and the reason is that they are formed by the atoms of the material always carrying themselves in the same pattern. This regular arrangement causes the regular shape of crystals, and also makes it possible for a crystal to conduct electricity. For any material to conduct electricity, it must be well supplied with particles smaller than atoms which have an electric charge, positive or negative, and these particles must be able to move freely through the material.

The regular arrangement for atoms in crystals provides plenty of paths between the atoms for the easy movement of these charged particles, so that crystals only need a supply of particles to become conductors. The materials we call metals are crystals which can release about one charged particle from each atom, so they conduct electricity pretty well, though not equally well. Insulators, on the other hand, simply don't have many charged particles lying around and many of them aren't crystals either, making it doubly difficult for them to conduct. In between these two extremes are the curious materials called semiconductors, which form crystals but are not well supplied with the charged particles that are needed to make them into conductors.

These are two ways in which we can supply these particles. One way is to heat the materials. This causes a few atoms to shed one of their electrons (negatively charged particles), leaving behind a gap in the arrangement of particles in the crystal which we call a hole. The hole behaves like a positively charged particle and can slip from one atom to another. Raising the temperature of a semiconductor, therefore, makes it conduct, but the electrons will slip back into place again when the material cools so the change is not permanent.

Dope Charge

A permanent change can be caused by doping. Doping is adding a small amount of impurity to a semiconductor material. We don't use any old impurity, but materials whose atoms will fit nicely into the arrangement of atoms in the crystal. Some of these materials which fit perfectly into place have one electron more than is needed in the crystal. That electron is released from each impure atom, allowing the crystal to conduct electricity by movement of these electrons. A crystal doped in this way is called N-type. We can also dope with a material which has fewer electrons than its neighbours in the crystal, creating a hole and making the crystal conduct by hole movement. A crystal doped in this way is called P-type. When a semiconductor is made into a conductor by doping, the change is permanent because there are always electrons or holes which don't fit and can't just snap together again (recombine).

This business of doping is quite something, because

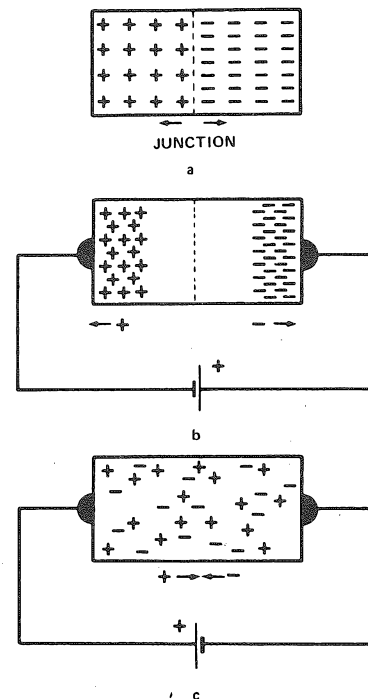


Fig. 4. When a junction is formed (a) the electrons and holes separate slightly at the junction. Reverse bias (b) makes the separation much greater so that the material can't conduct — there aren't any carriers. Forward bias (c) allows electrons and holes to cross the junction, making the material a conductor.

it allows us to do a bit of engineering on materials, creating crystals which can be fair conductors or good conductors, according to how much doping we use; or which are N-type or P-type according to what type of doping we use.

Attractive Likes

Now we've set the scene for learning why a diode works, and there's only one main point left. Charged particles, whatever their size, obey the laws of electrostatics. Of these laws, the important one for understanding the action of a diode is that two particles with the same sign of charge (two positives or two negatives) will repel each other, but particles with opposite signs (a positive and a negative) will attract each other. It's a simple enough law, but combined with what we now know about doping it's enough to explain what goes on inside a diode.

A diode is a single crystal with P-type doping on one end, or on one face, and N-type doping at the other end or face. Obviously, there's got to be a surface in the middle or thereabouts where these two types of doped material meet, and this surface is called the junction. The important thing about a junction is that it's somewhere inside a crystal with no break in the arrangement of the atoms. You can't make a junction by pressing a lump of P-type material up against a lump of N-type material — there's no chance that the rows of atoms would ever line up the way they do inside a crystal.

This arrangement is now a diode — a crystal with P-type material on one side of the junction and N-type material on the other. Remember what these terms mean — N-type material conducts because it has electrons free to move through the crystal. Because the crystal is in one piece, there's no reason why electrons or

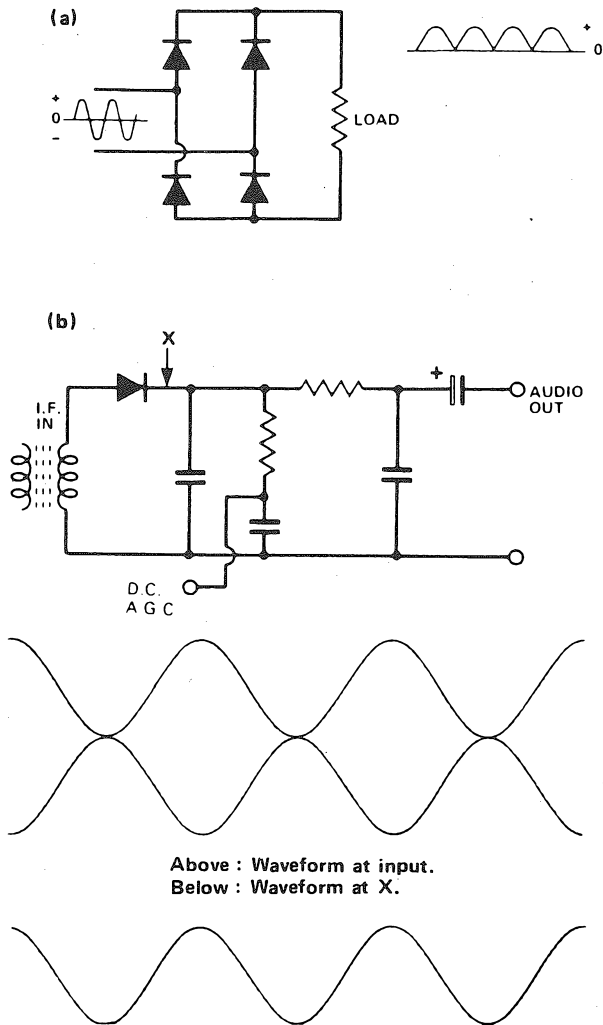


Fig. 5. Using diodes (a) for rectification (b) for radio signal detection. Both applications depend on the one-way flow of current through the diode.

holes should not move from one end of the crystal to the other, so the crystal can be made part of an electrical circuit.

Up The Junction

When the junction is formed, though, the free electrons of the N-type material at one side will be placed very close to the free holes in the P-type material on the other side and inevitably there's a bit of shuffling which ends up with some combination of electrons and holes. This leaves the junction without carriers and also causes the carriers to be pulled back a bit from the junction. The carriers are pulled back because the electrons removed from the N-type material leave a positive charge behind — originally there must be a positive charge for every electron — and the holes that are removed from the P-type material leave electrons (negatively charged) behind.

The effect of the remaining charges is to attract electrons and holes (carriers) away from the junction (Fig. 4a). The imbalance of charge also shows up as a voltage and this is what causes the OV5 or so we need before we can make a silicon junction conduct in the forward direction. The bit of crystal around the junction that has no free carriers is called the depletion layer and we'll look at it again when we discuss varicap diodes.

Minority Groups

The action of the diode in a circuit now becomes a bit easier to understand. When the diode is reverse biased, the polarity of the power supply (Fig. 4b) acts to attract carriers away from the junction, making the depletion layer wider. The electrons of the N-type material and holes of the P-type material simply don't cross the junction because they are pulled in the opposite direction. The only carriers that can cross are what are called minority carriers, holes which appear in the N-type material and electrons which appear in the P-type material. These minority carriers come from splitting bits off atoms in the crystal, using energy from the action of temperature or light. The higher the temperature of the diode the faster these minority carriers are formed. If we make the reverse voltage across the depletion layer high enough, the effect will be to accelerate these minority carriers to high speeds, so that they bang into atoms, knock more carriers off, and so cause the whole junction to become conducting. When that happens, the junction has 'broken down', the diode conducts and it can be damaged.

When the bias is in the forward direction (Fig. 4c) the carriers are attracted towards and across the junction. First of all, though, the voltage caused by the depletion process has to be overcome. Once the forward voltage has reached this amount, current starts to flow. Only a few of all the possible carriers cross the junction when the voltage is low, but raising the voltage even by a very small amount is enough to cause a great increase in the number of carriers crossing over the junction, so that the resistance of the junction becomes much less as the voltage and current are increased.

Shedding Light

This picture of what is happening inside a diode explains pretty well the action of signal or rectifier diodes which are used in the circuits such as those shown in Fig. 5. What about some of the other diodes that we use, like photodiodes, varicaps, LED's, and Zeners? Let's start with photodiodes. The main difference between a photodiode and an ordinary signal diode is that we deliberately put a photodiode into a transparent case so that light can reach the junction. Photodiodes are used in circuits where they are reverse-biased, with a fairly wide depletion layer. Now in darkness, the amount of current that can flow is only that caused by minority carriers — the few holes and electrons that are split off by the heat of the surroundings. Light, however, is a wave which, like all waves, carries energy. The energy of light falling on the depleted layer around the junction can cause lots more electrons and holes to be split off.

They're still minority carriers, but there's a lot more of them now, and so a layer current flows despite the

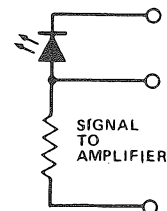


Fig. 6. Using a photo-diode as a light detector. The diode is reverse-biased, but will conduct slightly when light separates electrons from holes.

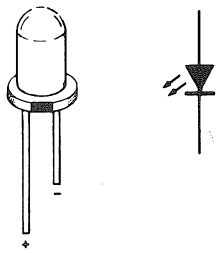


Fig. 7. The LED. When forward current flows, a glow of light is visible. Beware of reverse voltages — anything more than about 3V reverse will destroy the junction.

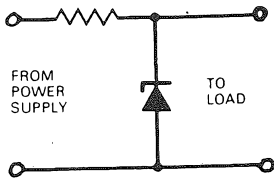


Fig. 8. The Zener diode used as a simple stabiliser. A load connected across the diode can draw current by reducing the current through the diode. Providing the diode current doesn't drop below about 2 mA, the voltage across the diode will remain constant.

reverse bias. Typically, the reverse current can change from around 0.1 μA in darkness to 100 μA in the light of a desk lamp. If the diode is forward biased, the change caused by light is hardly noticeable.

Together

The LED has an action which is just the reverse of that of the photodiode. Instead of light falling on the junction and causing electrons and holes to split off, as happens in the photodiode, the LED depends on electrons and holes coming together again and giving out light. You can imagine these two processes more clearly when you think of separating two strong magnets. The force which holds them together means that you have difficulty separating them — you have to do some work to separate them. You can get that work back again when the magnets attract each other back; you could even

make the magnetic force do something useful, like picking up a weight.

LED Light

LED's are made from semiconductors (such as Gallium phosphide) which are not heavily doped and don't conduct very well. Something like 2 V is needed across the junction of a typical LED to get current flowing and the movement of holes and electrons causes collisions which separate off more holes and electrons. On their way across the junction in opposite directions, holes and electrons collide — and release the energy it took to split them apart in the first place. The amount of energy is the same as that of a light wave and since the material is transparent a light wave is what we get. The colour of the light wave is decided by how much energy is released. Low energy gives red light, or the invisible infra-red. Higher energy gives yellow, green, blue light (in order of increasing energy), until we reach the invisible ultraviolet radiation. The amount of energy is fixed by the material that is used as a semiconductor, though, and we can't alter it noticeably by changing the voltage or current.

Avalanche

Zener diodes make use of the reverse breakdown which has already been described. Oddly enough, two effects cause this reverse breakdown, Zener effect and avalanche effect. The avalanche effect is the one we've described, in which minority carriers are accelerated so much by the reverse bias that they collide with atoms and split electrons and holes apart. This creates more carriers, which are in turn accelerated, splitting off yet more until the whole junction becomes conducting. The avalanche effect occurs mainly in lightly doped material, at reverse voltages of 6 V or more. The other effect, Zener effect (named after Carl Zener who discovered it) takes place in heavily doped materials, mainly when the reverse voltage is less than 6 V. Because of the large number of electrons and holes which are present, the depletion layer is very thin and it's comparatively easy for a carrier to shoot straight across. Diodes which made use of either or both of these effects are called

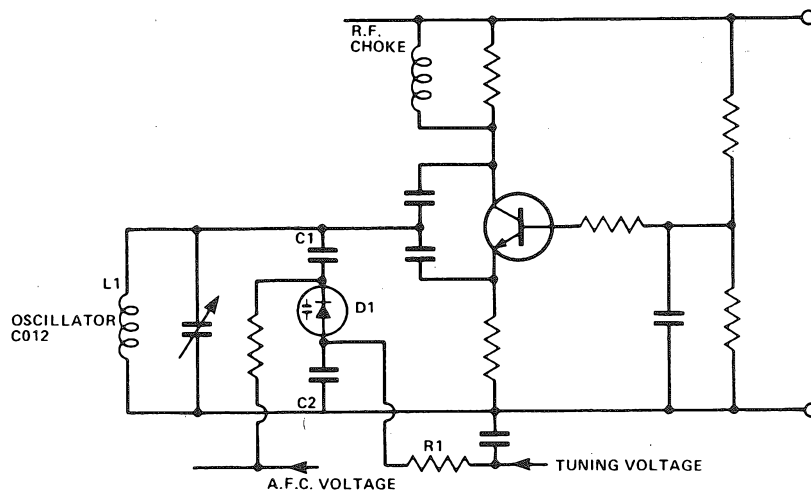


Fig. 9. The varicap diode, D1, is in series with C1 and C2, and is part of the tuning capacitance for L1. Since the diode capacitance is varied by the control voltage from R1, tuning can be carried out by altering this DC voltage.

Zener diodes, and we use them to stabilise voltage. The breakdown, particularly when it is caused by avalanche effect, takes place at a precise value of voltage, so that a Zener diode wired in the circuit of Fig. 8 will have an almost constant voltage across it, even if the current through it varies considerably.

Incidentally, avalanche effect has a positive temperature coefficient, which means that the voltage across the junction increases as the temperature is increased. Zener effect, by contrast, has a negative temperature coefficient, meaning that the voltage across the junction decreases as the temperature is increased. At voltages around 5V6, both effects take place, which means that the voltage is hardly affected by temperature. For this reason, 5V6 zener diodes are often specified rather than any other voltage.

Varicaps

Finally, among the diodes that are particularly useful, varicap diodes make use of the width of the depletion layer. The depletion layer, remember, is the part of the crystal around the junction which has had its carriers removed. The greater the reverse bias applied to the diode, the greater the attraction of carriers away from the junction and so the greater the width of the depletion layer.

Now a depletion layer is a chunk of insulating material which is sandwiched between two bits of conductor — the P and N materials. This is just the arrangement we know as a capacitor — an insulator between two con-

ductors — so that the reverse-biased diode has a capacitance. It's a variable capacitance, though, because the width of the insulator — the depletion layer — can be varied by changing the bias voltage. Like any other variable capacitor, the capacitance value is greatest when the insulating layer is very thin, and the capacitance value is least when the insulating layer is thick. Now the diode has a thick depletion layer when the reverse voltage is large, so that its capacitance is low; but when the reverse bias is small, the depletion layer is thin and the capacitance is large.

Varicap diodes solve an awkward problem — how to tune radio circuits without having any moving parts. A varicap diode in the oscillator circuit (Fig. 9) arranged in series with a fixed capacitor so that it is only part of the tuning capacitance, has no DC connection to the oscillating circuit and can have its capacitance varied by a voltage supplied from a potentiometer. The potentiometer doesn't have to be anywhere near the tuned circuits, so long as the connecting wires are well decoupled and the tuned circuits can be sealed inside a can, undisturbed by any movements.

That's dealt with the most common diodes, though there are dozens of types we haven't mentioned, ranging from the diodes which generate microwave signals to the breakdown diodes we use in thyristor firing circuits. Once you've grasped the basic principles, though, there aren't many surprises left, and you are better able to understand how to make efficient use of these indispensable components.

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DESIGNING OSCILLATORS

One of the problems in electronics is stopping amplifiers from oscillating, another problem is getting oscillators to oscillate . . . Tim Orr explains.

An oscillator is basically an amplifier with positive feedback applied around it. The feedback must be AC coupled otherwise a DC latch up condition would occur. Having got some sort of oscillation, one of two things can happen. The oscillation can build up in amplitude until clipping occurs due to the power supply voltage levels. At this point a stable, but truncated waveform will be generated. Alternatively if the gain of the amplifier is too low the oscillation will die away. To produce a pure sinusoidal oscillation the level of the signal in the system must be accurately controlled. There must be some amplitude limiting or automatic gain control such that when the peak signal level tries to exceed a reference voltage, the amplifiers gain is reduced. This is in fact what limiting does. To maintain stable oscillation, the overall gain of the system must be exactly unity. Any less and the oscillations will never start. If the gain is more than unity, the oscillations will occur, but amplitude limiting will cause gross distortion.

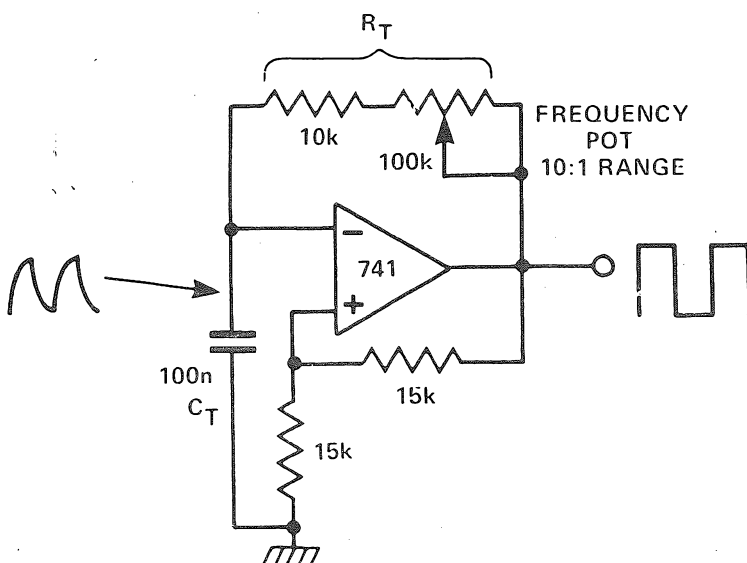
A very common method for stabilising the oscillations, which is often used in Wein bridge oscillators, is to employ a very sensitive thermistor as an AGC. However, the thermal time constant of this component often produces an annoying amplitude bounce which occurs

when changing to a new frequency.

Other methods are diode limiters (which tend to cause large amounts of distortion) and FET AGC circuits. The latter method can be used to generate super low distortion sinusoids by allowing the system gain to stabilise over tens of seconds.

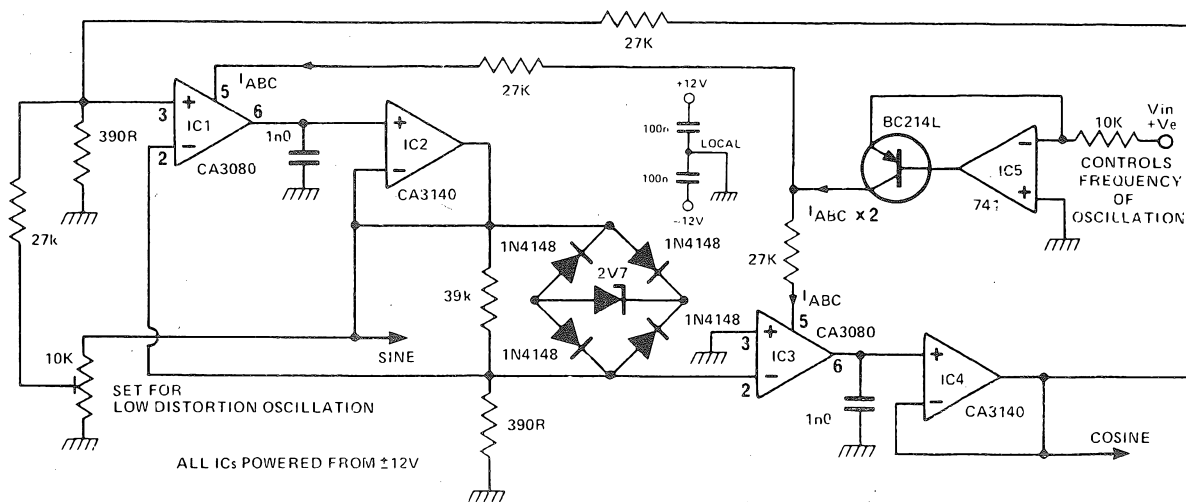
The oscillation frequency is mainly determined by the feedback around the amplifier. By making the feedback a reactive network, the phase of the feedback will vary as a function of frequency. Oscillations can only occur when the feedback is positive and thus the phase response of the feedback will determine the frequency of oscillation, assuming that the overall gain at this frequency is at least unity. By varying the phase response of the feedback, the oscillation frequency may be altered.

An oscillator should be thought of as being a circuit which continuously generates a waveform, no matter what the shape of the waveform. There are very many circuit techniques for generating these signals which range from relaxation oscillators to piece wise approximations using square waves. Some of these methods will now be illustrated.



Manually Controlled Oscillator

In this circuit there are two feedback paths around an op-amp. One is positive DC feedback which forms a Schmitt trigger, the other is a CR timing network. Imagine that the output voltage is +10V. The voltage at the non-inverting terminal is +5V. The voltage at the inverting terminal is a rising voltage with a time constant of $C_T R_T$. When this voltage exceeds +5V, the op amp's output will go low and the Schmitt trigger action will make it snap into its negative state. Now the output is -10V and the voltage at the inverting terminal falls with the same time constant as before. By changing this time constant with a variable resistor a variable frequency oscillation may be produced.



Dual Integrator Quadrature VCO

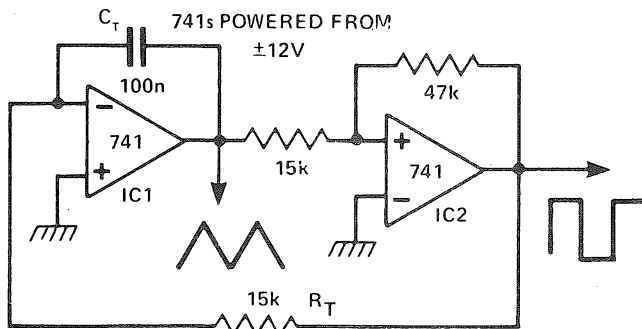
This is a sinusoidal oscillator which uses frequency dependent feedback and zener diode amplitude limiting. IC1,2,3&4 form a dual integrator circuit which is an analogue model of a second order differential equation! There is some positive feedback around IC1,2 which is analogous to having a zero damping factor in the equation. This means that the oscillations will build up. The positive feedback is controlled by the 10k preset. IC1,3 are integrators and IC2 and IC4 are voltage followers with high input impedance. The phase shift produced by an integrator is 90° so there is no overall feedback around the loop (IC1 is non-inverting, IC2 inverts). Thus we have all the conditions for oscillation, and in fact oscillations will occur when the preset is adjusted to give the correct phase shift around the IC1,2 stage. Amplitude limiting is produced by the 2V7 zener inside the diode bridge. By placing it inside the bridge the same diode is used for both positive and negative signals and the limiting is symmetrical. The integrators are two quadrant multipliers (CA3080s), so the gain of the loop can be controlled by the current I_{ABC} . In the solution of this second order differential equation, the gain

of the loop is proportional to the resonant frequency. Thus, by varying I_{ABC} or rather by varying V_{IN} , the frequency of oscillation may be altered.

As the integrators produce a 90° phase shift, the two sinusoid outputs are in phase quadrature, i.e. one is a sine wave, the other a cosine wave. The cosine output is lower in distortion than the sine wave, because the amplitude limiting (and hence the distortion) is produced at the IC1,2 stage.

The second stage (IC3,4), acts as a filter and hence produces a purer sinusoid. Using this circuit a 1000 to 1 continuous frequency sweep can be obtained. However, the inaccuracies in the CA3080's will cause some amplitude variations and it may be necessary to set the positive feedback a bit high (and hence attract more distortion), to maintain stable amplitude limiting over the sweep range. This circuit is an oscillating filter and if you turn down the positive feedback and inject a small signal through a 100k resistor into IC1 pin 3, a bandpass and low pass response is obtained from the sine and cosine outputs respectively.

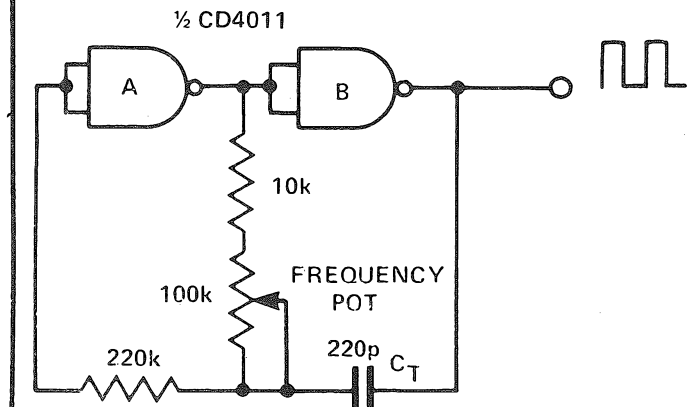
Simple Triangle Square Wave Oscillator



This circuit generates simultaneously a triangle and a square waveform. The triangle could be 'bent' by a diode function generator to produce a sine wave. The circuit is always self starting and has no latch up problems. IC1 is an integrator with a slew rate determined by C_T and R_T and IC2 is a Schmitt trigger. The output of IC1 ramps up and down between the hysteresis levels of the Schmitt, the output of which drives the integrator. By making R_T variable it is possible to alter the operating frequency over a 100 to 1 range. Three resistors, one capacitor and a dual op amp is all that is needed to make a versatile triangle squarewave oscillator with a possible frequency range of 0.1Hz to 100kHz.

CMOS Oscillator

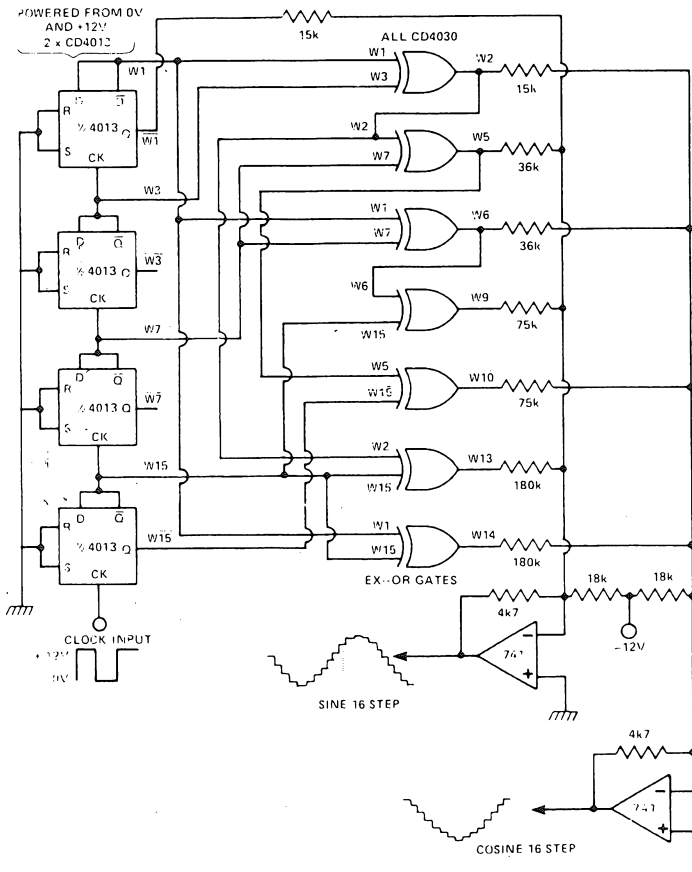
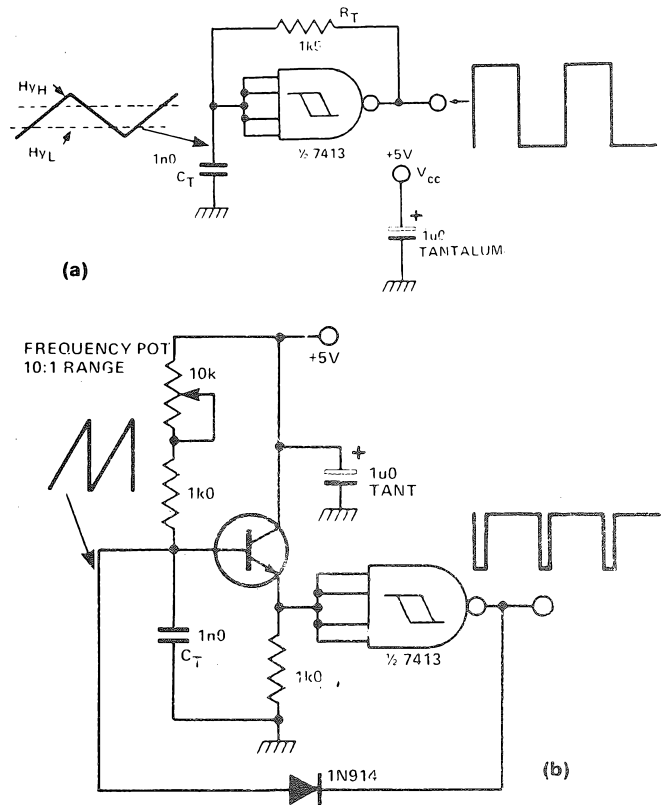
Two CMOS gates can be used to produce a simple oscillator. Imagine that output B is high. Then the input to A is also high due to it being coupled via the capacitor C_T to output B. Thus output A is low, input B is low and output B is high, which is as we would expect. However, capacitor C_T is being discharged via the 100k pot and 10k resistor to a logic 0. When this voltage reaches the crossover point for A, output A goes high, and thus output B goes low. Now the capacitor is charged up to a logic 1. Thus the process repeats itself. Varying the 100k pot changes the discharge rate of C_T and hence the frequency. A square wave output is generated. The maximum frequency using CMOS is limited to 2MHz.



TTL Oscillator

A simple relaxation oscillator can be made using a TTL Schmitt trigger. The circuit 'a' is the most simple version that can be produced. Imagine that the output is high. Capacitor C_T is charged up via R_T . When the upper hysteresis level (H_{yh}) is reached, the output goes low. C_T is now discharged until the low hysteresis level (H_{yl}) is reached whereupon the output goes high. Thus the oscillator generates a square wave, with an uneven mark to space ratio, due to the input current requirements of the 7413. The frequency can be set at any value up to several megahertz by varying C_T and R_T . C_T can be an electrolytic but R_T must not be more than about 1k5 or it will not be able to pull down the Schmitt trigger inputs. (If you use a CMOS Schmitt this does not apply). The output is a nice fast squarewave capable of directly driving several TTL loads. One problem to be encountered is frequency jitter. When the input is very near to a hysteresis level, noise in the system may cause the oscillator to prematurely trigger, thus making that period slightly shorter and producing a noise induced frequency jitter. Also using two Schmitt triggers from the same IC is sure to cause interaction and thus jitter. To reduce power supply noise effects the IC should be decoupled with a 1uF tantalum capacitor actually at the V_{cc} and GND pins of the package.

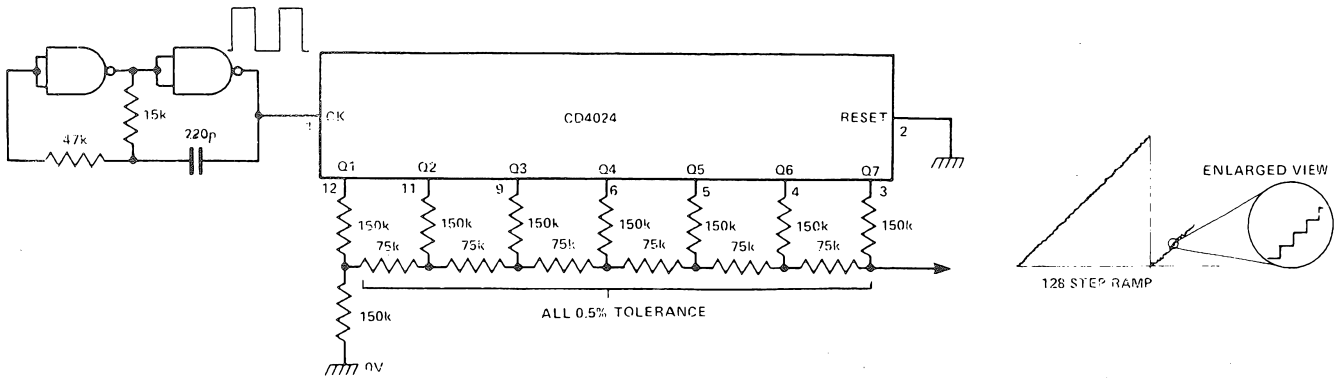
Diagram 'b' shows the same oscillator, but with a 10 to 1 manual control of frequency. The timing capacitor is charged up by the 10k pot and the 1k resistor. This voltage is then buffered by the emitter follower and fed to the Schmitt trigger. When the upper hysteresis level is reached the output of the Schmitt goes low and the capacitor is rapidly discharged via the diode until the lower level is reached. The process then repeats itself. As the discharge period is so fast, (it can be as short as a few hundred nano seconds) the period can be thought of as being determined by the charging time, which is controlled by the 10k pot.



Walsh Function Generator

The mathematician, Fourier, said that any repeating waveform could be made up out of harmonic components. These components are sinusoids which are integrally related to the fundamental period of the waveform in question. This is a convenient conceptual approach, but as a way of practically synthesising waveforms it is not on. You would have to generate a whole series of harmonically related sinewaves which might prove a little difficult. However, a man called Walsh said that you could do the same thing as Fourier, but with square waves. So, instead of using sinusoidal Fourier sets, we can use square wave Walsh functions to synthesise waveforms. There are various techniques for calculating the Walsh function co-efficients for generating particular waveforms but these are beyond the scope of an article such as this. The diagram shows the circuit for generating sine and cosine waveforms using 16 steps. Walsh functions are orthogonal functions, just as sine and cosine are orthogonal, and so the generation of these two waveforms is relatively simple using this technique. The 4013 dividers and the exclusive OR gates generate the Walsh functions, which in turn are converted into analogue waveforms by use of the correctly weighted resistor networks. Note that you only need 4 resistors to generate a 16 step sinewave approximation.

The resultant outputs can be easily filtered by fixed or tracking filters to produce pure sinusoids. The output frequency is 1/16th of the input clock frequency. The clock can be stopped and the outputs will remain fixed, try that with analogue techniques!

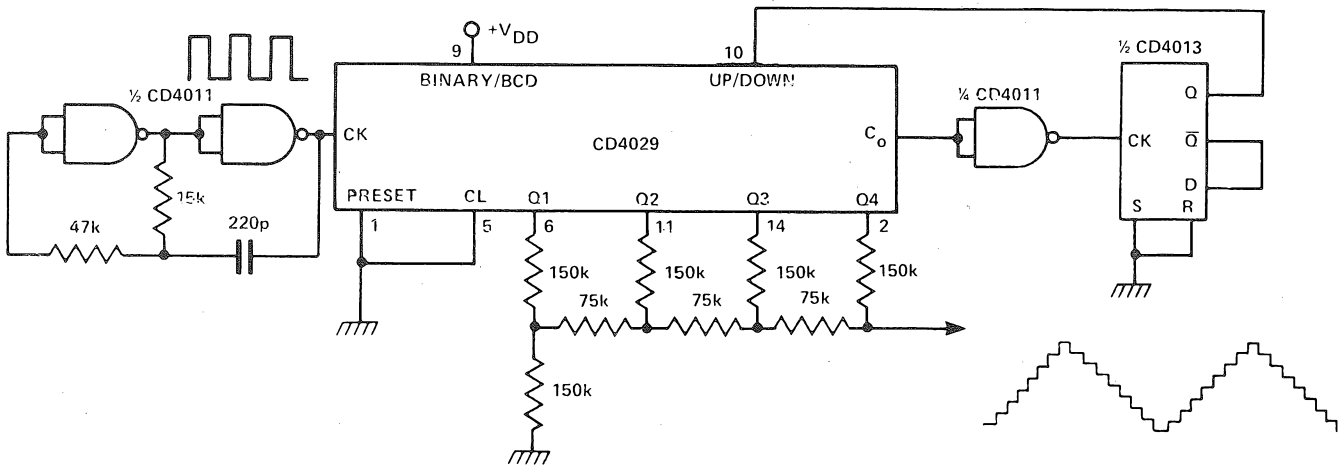


R-2R Staircase Generator

Waveforms can be constructed by building them up out of separate elements. In this case a linear ramp waveform is generated out of 128 steps. The CD4024 is a seven stage binary counter. It is being driven from a CMOS clock oscillator similar to that already described.

The Q1 to 7 outputs divide this clock frequency by

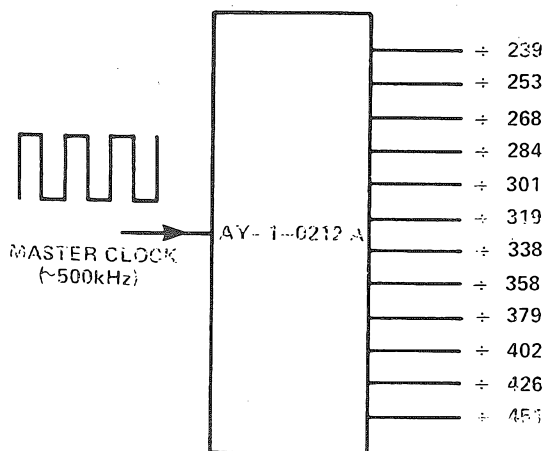
2,4,8,16,32,64 and 128 respectively and the divided outputs are then fed into an R,2R ladder network. This is in fact a Digital to Analogue Converter (DAC) and as the counter is merely counting up, then the converter will generate a linearly rising waveform made out of 128 steps. When the counter overflows, the ramp waveform resets and the process repeats itself.



R-2R Triangle Generator

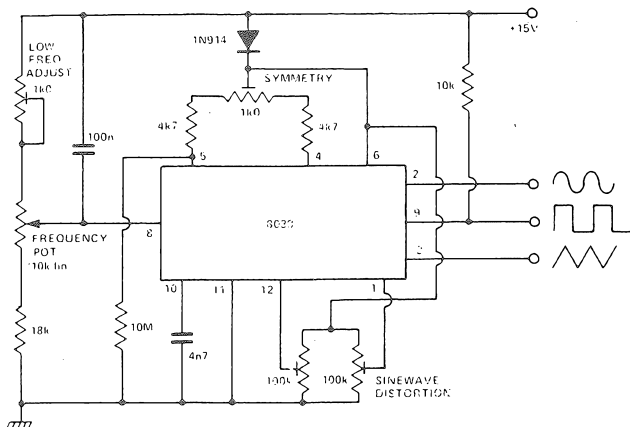
This circuit is similar to the previous except an up/down counter is included. A clock signal is applied to the 4029 counter. When it has counted 16 clocks a Carry signal is generated. This clocks a D type flip-flop (4013), which changes state and reverses the up

down mode of the 4029. Thus the circuit counts up, down, up, etc. The counting is converted via an R,2R ladder into an analogue output, a triangle waveform made up out of several steps.



Master Tone Generator

If you have ever made an electric organ, piano or string machine you would have had to produce the top twelve notes for the top octave by some means or other. More expensive organs might use 12 master oscillators which would be tuned to the top twelve semitones on the keyboard. This gives a nice free phase quality to the sound. The notes in the octaves below are made by using binary dividers and filtering. Very expensive organs would use an oscillator per note. This allows every note to be individually tuned and produces a very good sound quality. However, there is an easy way of producing the semitones and this is with a master tone generator chip. This is a pre-programmed divider having one input and twelve or thirteen outputs. A high frequency master clock is put into the chip which is divided by numbers ranging from 239 to 451. These divisions produce the semitone outputs. Thus, by using one master oscillator and one master tone generator a lot of the work of making an organ is removed. It is possible to produce more accurate intervals using 12 oscillators, but the speed and efficiency of the chip usually wins in the lower price end of the market.

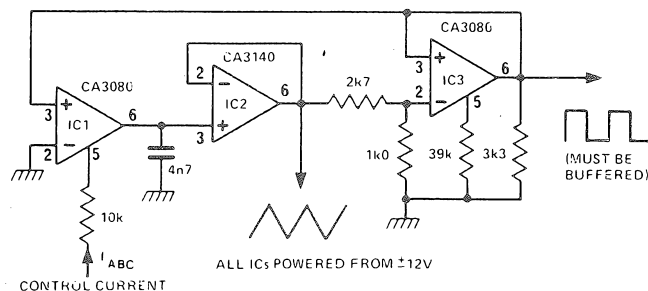


8038 Function Generator

There are several ICs available which perform some sort of oscillator function. One such is the Intersil 8038 which is a VCO with sine, triangle and squarewave outputs. The basic oscillator is a triangle squarewave device with a function generator to produce the sinewave. The frequency is voltage controllable but is not a linear function. The triangle symmetry and hence sinewave distortion are adjustable with a preset but change when the frequency is altered. Operation up to 1MHz is possible.

Triangle Squarewave ICO Using CA3080's

This circuit is very similar to that of the simple triangle/square oscillator, except that the operating frequency is controlled by a current I_{ABC} (ICO stands for current controlled oscillator, as opposed to VCO, voltage controlled oscillator). Using this circuit, a sweep range of 10,000 to 1 is possible (for I_{ABC} 500 μ A to 50nA). The CA3080 is a two quadrant multiplier and the CA3140 is a MOS FET op-amp. IC1 is used as an integrator. IC2 is a high input impedance voltage follower and IC3 is a Schmitt trigger. The CA3080 has a current output which in the case of IC1 is used to charge up a capacitor. The voltage on this capacitor is buffered by the CA3140 and fed into the Schmitt IC3. The CA3080 (IC3) forms a very fast Schmitt trigger but as it has a current output, it cannot be loaded in any way without affecting the operating frequency. The output of the Schmitt is used to make the integrator inverting or non-inverting. Thus the operation is as follows. The integrator ramps upward until the positive hysteresis level is reached. The Schmitt flips over, the integrator then ramps downwards until the negative hysteresis level is reached. The Schmitt flips back and the process is

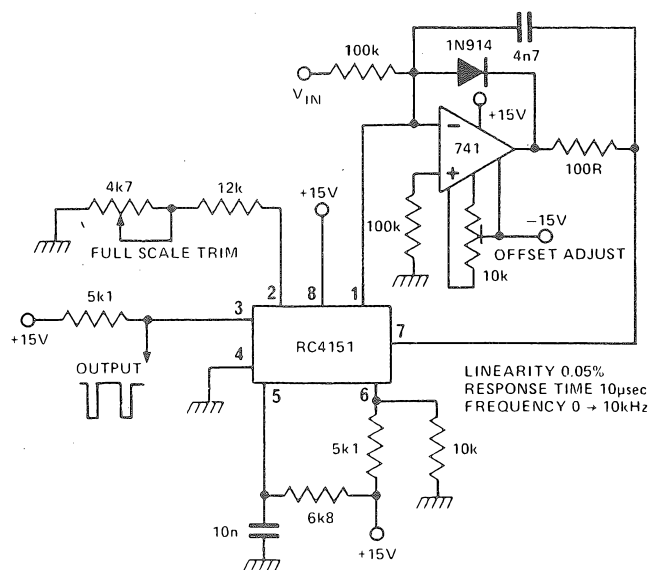


repeated. The ramp rate is determined by the size of the current I_{ABC} and therefore I_{ABC} is linearly proportional to the oscillation frequency. At very low currents the triangle waveform may become very asymmetrical. This is due to current mirror mismatches inside IC1 and this device may have to be specially selected for continuous symmetry.

Precision Voltage Controlled Oscillator

The RC 4151 is a precision voltage to frequency converter. It generates a pulse train output which is linearly proportional to the input voltage. The linearity for the circuit shown is 0.05%. The IC compares the input voltage with an internally generated one. It dumps controlled pulses of charge into a parallel RC network and compares this generated voltage with the input. If the input is greater it puts more pulses of charge into the RC network until the two are balanced. To get a larger sustained voltage in the RC network the frequency of the pulses must be increased. Thus the frequency of the pulses generated is made to be proportional to the input voltage.

The output is a pulse waveform and is intended to drive some sort of counting system, the chip being used as simple analogue to digital converter. It can also be used as a frequency to voltage converter. A maximum frequency of 10kHz has to be observed.



LINEARITY 0.05%
RESPONSE TIME 10 μ sec
FREQUENCY 0 → 10kHz

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Introduction

Since ETI was originally launched in Britain in April 1972 we have published nearly 5000 pages of editorial matter, much of this has been devoted to electronic projects of course. In our history we have republished the best of these projects in Top Projects No. 1, No. 2, we have also reprinted the Tech-Tips and other circuits in Circuits Book No. 1 and No. 2 and our beginners series Electronics It's Easy is now on its third reprint. We have not however, until now, brought together the best features dealing with electronic design. Nearly

1500 pages have been devoted to this in recent years and this special is the first of a series of three which will bring together the best 250-300 of these.

Although some features are from quite early issues they have only been included because we feel that age has not withered them.

We hope that the contents help the reader understand at least some of the aspects of electronic circuit design.

Halvor Moorshead
Editorial Director - ETI

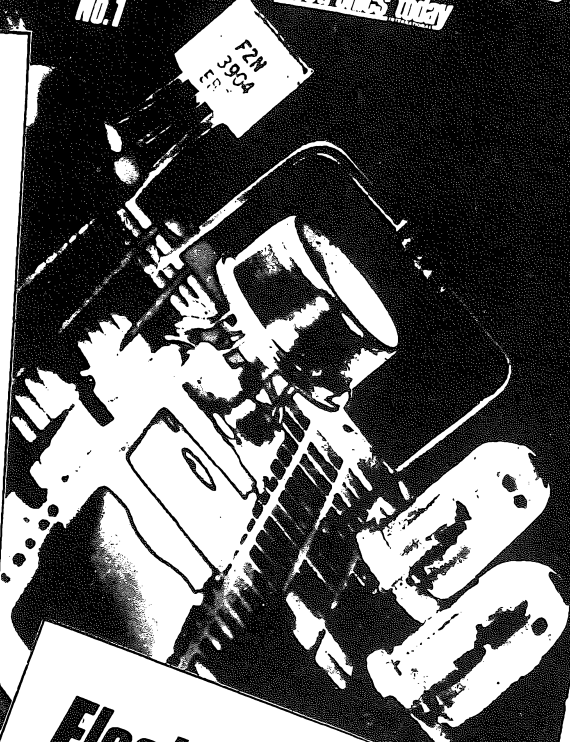
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Introduction

This special publication from ETI magazine is the second in a series of three, the first was published in late autumn 1979.

The series comprises the best 250-300 pages selected from well in excess of 1500 published in ETI on aspects of circuit design during its history.

The articles included in this series are intended to give the reader an insight into how circuits work, how they

are designed and to provide some fairly conventional "building blocks". Keeping track of current design techniques is not easy due to the rapid introduction of new devices and new technologies we hope this publication helps to cover the problem.

Halvor Moorshead
Editorial Director - ETI

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