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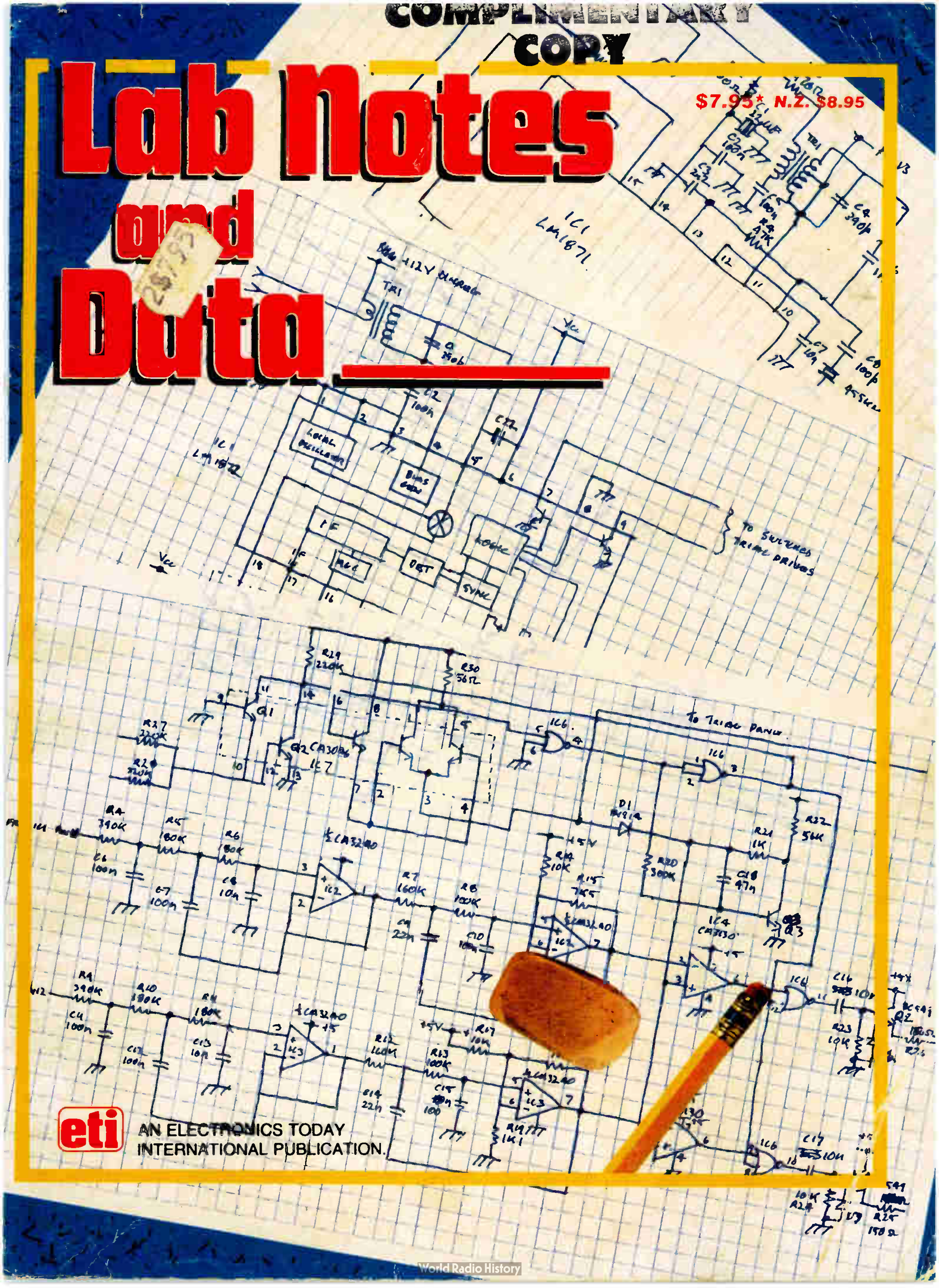
Lab Notes and Data



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Lab Notes and Data

LAB NOTES

DATA SHEETS

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Using 3080 op-amps

A control current is used to vary the gain of this op-amp.

by Tim Orr

THE CA3080 IS KNOWN as an *operational transconductance amplifier* (OTA). This is a type of op-amp, the gain of which can be varied by means of a control current, (I_{ABC}). The device has a differential input, a control input known as the 'amplifier bias input' and a current output. It differs in many respects from conventional op-amps and it is these differences that can be used to realise many useful circuit blocks.

Voltage controlled amplifier

The CA3080 can be used as a gain controlling device. A useful circuit is shown in Figure 1. The input signal is attenuated by R1, R2 such that a 20 mV peak-to-peak signal is applied to the input terminals. If this voltage is much larger, then significant distortion will occur at the output. In fact, this distortion is put to good use in the triangle-to-sinewave converter. (Figure 3, but we're jumping the gun).

The gain of the circuit is controlled by the magnitude of the current I_{ABC}. This current flows into the CA3080 at pin 5, which is held at one diode voltage drop above the −Vcc rail. If you connect pin 5 to 0 V, then this diode will get zapped (and so will the IC!). The maximum value of I_{ABC} permitted is 1 mA and the device is 'linear' over four decades of this current. That is, the gain of the CA 3080 is 'linearly' proportional to the magnitude of the I_{ABC} current over a range of 0.1 uA to 1 mA. Thus, by controlling I_{ABC}, we can control the signal level at the output.

The output is a current output which has to be 'dumped' into a resistive load (R5) to produce a voltage output. The output impedance seen at IC1 pin 6 is 10k (R5), but this is 'unloaded' by the voltage follower (IC2) to produce a low output impedance.

The circuit involving IC3 is a precision voltage-to-current converter and this can be used to generate I_{ABC}. When Vin (control) is positive, it linearly controls the gain of the circuit. When it is negative, I_{ABC} is zero and so the gain is zero.

This type of circuit is known by several names. It is a *voltage controlled amplifier*, (VCA), or an *amplitude modulator*, or a *two quadrant multiplier*.

One problem that occurs with the CA3080 is that of the 'input offset voltage'. This is a small voltage diffe-

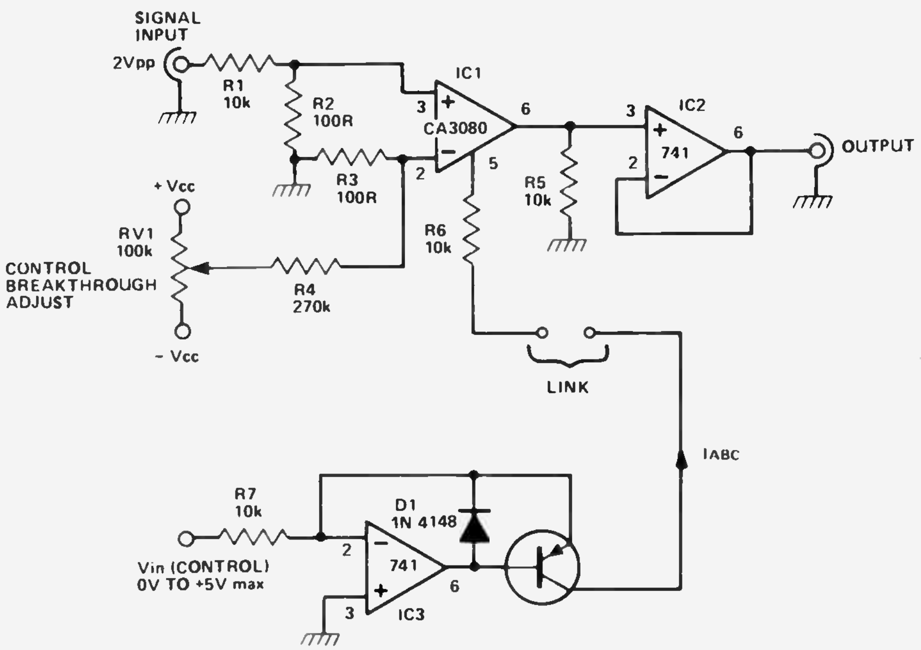


Figure 1. **A voltage controlled amplifier. Gain is varied by varying RV1. You can modulate a signal passing through the amplifier by joining the 'link' and applying a modulating signal to the input of IC3 (at R7). This sort of circuit is also known as a 'two quadrant multiplier'.**

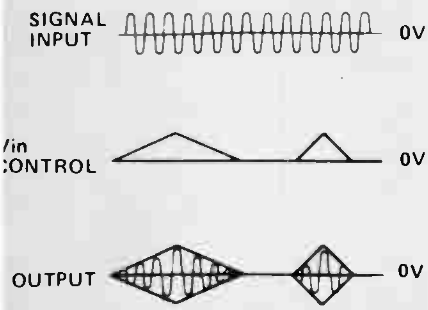


Figure 2. **Illustrating the operation of the voltage controlled amplifier shown in Figure 1.**

ence, or 'offset', between its input terminals. When there is no signal input and the control input is varied, a voltage similar to the control input will appear at the output. By adjusting RV1 it is possible to null out most of this control breakthrough.

The effect of modulating Vin (control) is illustrated in Figure 2.

Triangle to sinewave converter

By overloading the input of a CA3080 it is possible to produce a 'sinusoidal' transfer function. That is, if a triangle waveform of the correct magnitude is applied to the CA3080 input, the output will be distorted in such a way as to produce a sinewave approximation.

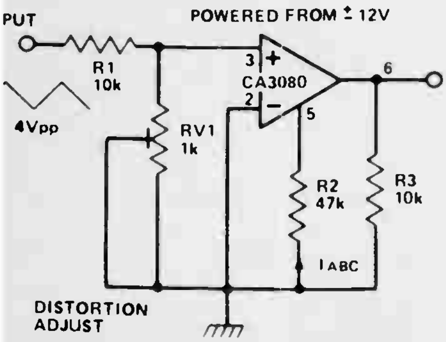


Figure 3. **This circuit will convert a triangle wave to a sinewave with a resultant distortion of around 1.8%.**

In the circuit shown (Figure 3), RV1 is adjusted so that the output waveform resembles a sinewave. I tested this circuit using an automatic distortion analyser and found the sinewave distortion to be only 1.8%, mostly third harmonic distortion which, for such a simple arrangement, seems very reasonable indeed. This could be used to produce a sinewave output from a triangle/square wave oscillator.

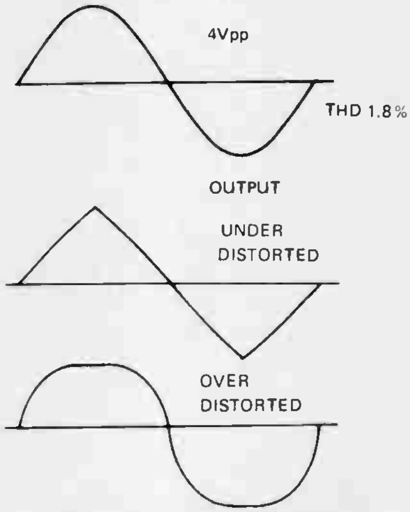


Figure 4. **The output of the Figure 3 circuit should be adjusted (by RV1) to produce the waveform shown at top.**

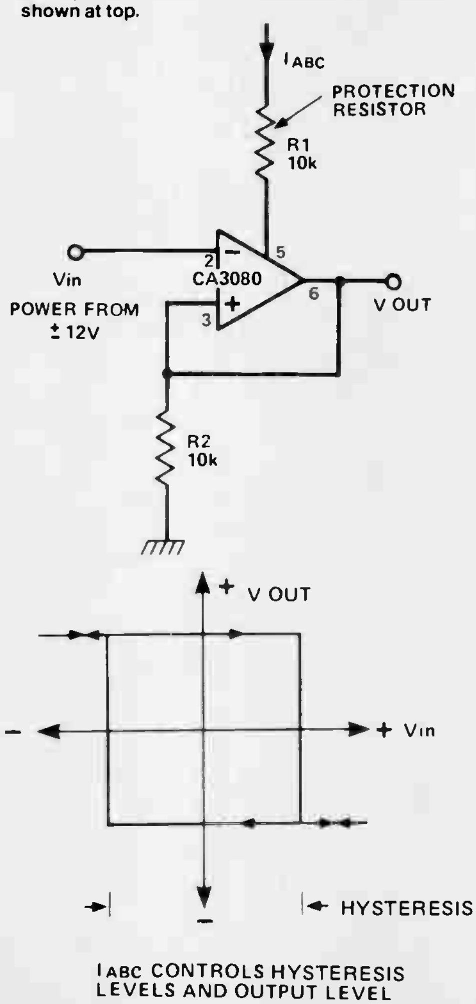


Figure 6. **This sort of Schmitt trigger is not only simple but you can specify the hysteresis levels as well!**

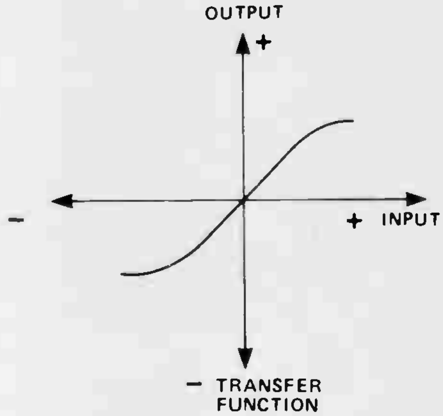


Figure 5. **Transfer function of the Figure 3 circuit.**

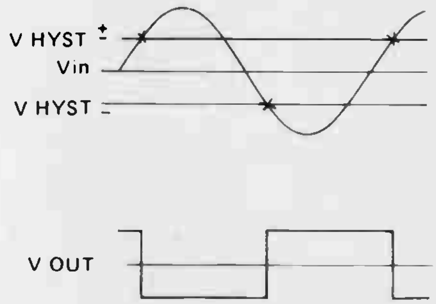


Figure 7. **How the Schmitt trigger of Figure 6 works.**

The result of varying RV1 is illustrated in Figure 4 and the transfer function of the circuit is shown in Figure 5.

Schmitt trigger

Most Schmitt trigger circuits prove to be very complicated when it comes to calculating the hysteresis levels. However, by using the CA3080 these calculations are rendered trivial, plus there is the added bonus of fast operation. The hysteresis levels are calculated from the simple equation,

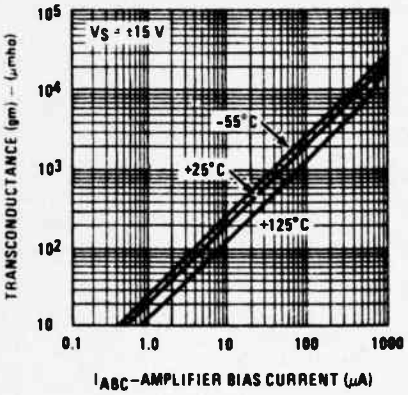
$$V_{HYST} = +/- (I_{ABC} \times R2)$$

The output squarewave level is in fact equal in magnitude to the hysteresis levels. The circuit operation is as follows (referring to Figure 7):

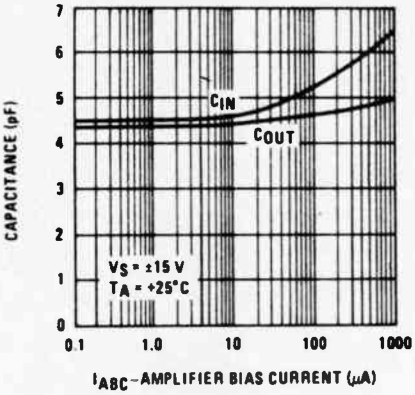
Imagine the output voltage is high. The output voltage will then be equal to ($R2 \times I_{ABC}$) which we will call $+V_{HYST}$. If V_{IN} becomes more positive than $+V_{HYST}$, the output will start to move in a negative direction, which will increase the voltage between the input terminals which will further accelerate the speed of the output movement. This ▶

SELECTED DATA ON THE 3080

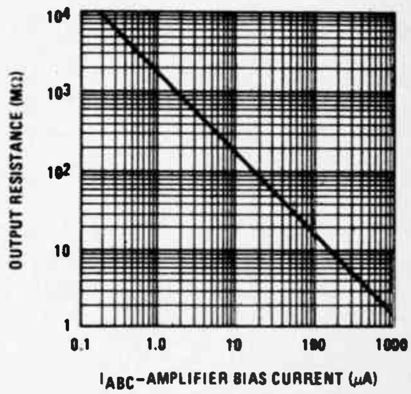
Transconductance



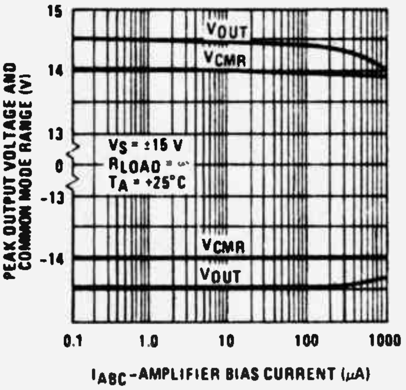
Input and Output Capacitance



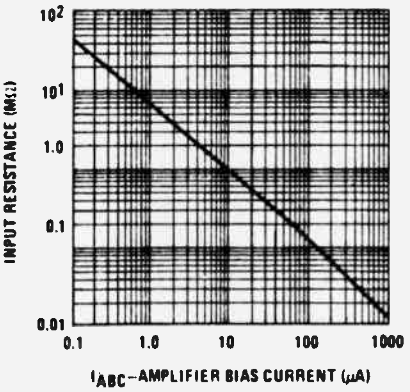
Output Resistance



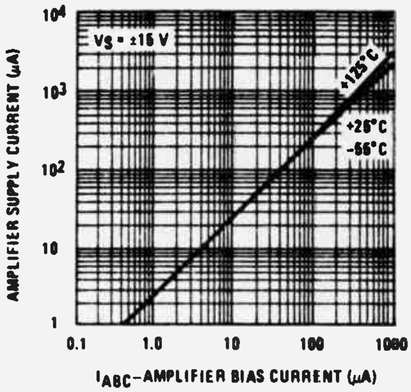
Peak Output Voltage and Common Mode Range



Input Resistance



Amplifier Supply Current



General description

The 3080 is a programmable transconductance block intended to fulfil a wide variety of variable gain applications. The 3080 has differential inputs and high impedance push-pull outputs. The device has high input impedance and its transconductance (gm) is directly proportional to the amplifier bias current (I_{ABC}).

High slew rate together with programmable gain make the 3080 an ideal choice for variable gain applications such as sample and hold, multiplexing, filtering, and multiplying.

Features

- Slew rate (unity gain compensated): 50 V/us
- Fully adjustable gain: 0 to gm R_L limit
- Extended gm linearity
- Flexible supply voltage range: +/−2 V to +/−18 V
- Adjustable power consumption

Absolute maximum ratings

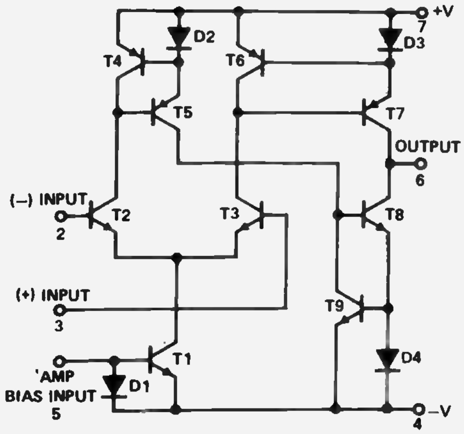
Supply Voltage 3080	+/−18 V
Power Dissipation	250 mW
Differential Input Voltage	+/−5 V
Amplifier Bias Current (I_{ABC})	2 mA
DC Input Voltage	$+/V_s$ to $-V_s$
Output Short Circuit Duration	Indefinite

Electrical characteristics, 3080 (Note 1).

Parameter	Conditions	Min.	Typ.	Max.	Units
Forward Transconductance (gm)		6700	9600	13000	umho
	Over Specified Temp. Range	5400			umho
Peak Output Current	R_L = 0, I_{ABC} = 5uA		5		uA
	R_L = 0	350	500	650	uA
	R_L = 0 Over Specified Temp.Range	300			uA
Peak Output Voltage					
Positive	R_L = , 5uA I_{ABC} 500uA	+12	+14.2		V
Negative	R_L = , 5uA I_{ABC} 500uA	−12	−14.4		V
Amplifier Supply Current			1.1		mA
Common Mode Rejection Ratio		80	110		dB
Common Mode Range		+/−12	+/−14		V
Input Resistance		10	26		k
Open Loop Bandwidth			2		MHz
Slew Rate	Unity Gain Compensated		50		V/us

Note 1: These specifications apply for V_S = +/−15 V and T_A = 25°C, amplifier bias current (I_{ABC}) = 500uA, unless otherwise specified.

Internal circuit of the 3080



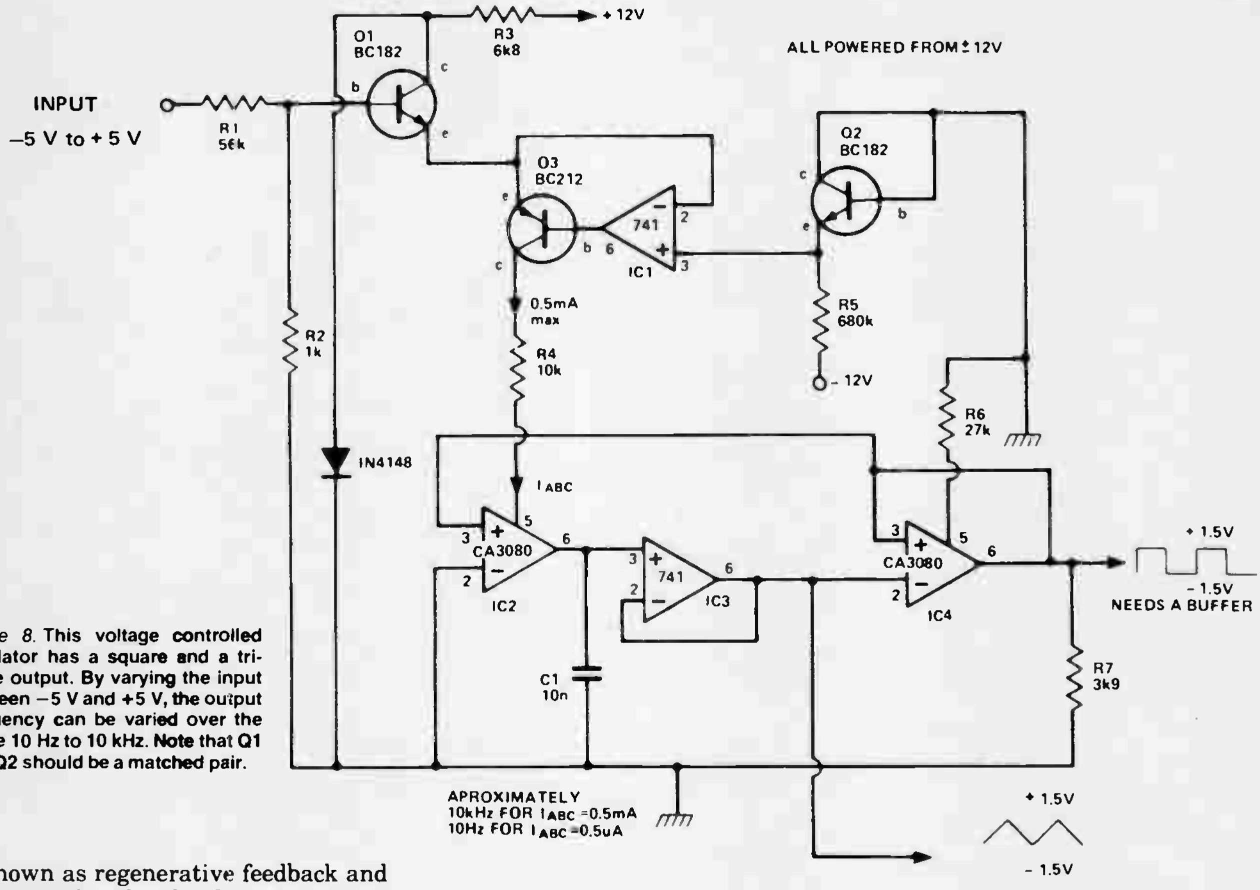


Figure 8. **This voltage controlled oscillator has a square and a triangle output. By varying the input between −5 V and +5 V, the output frequency can be varied over the range 10 Hz to 10 kHz. Note that Q1 and Q2 should be a matched pair.**

is known as regenerative feedback and is responsible for the Schmitt trigger action. The output snaps into a negative state at a voltage equal to $-(R2 \times I_{ABC})$ which is designated as $-V_{HYST}$. Only when V_{IN} becomes more negative than $-V_{HYST}$ will the output change back to the $+V_{HYST}$ state.

The Schmitt trigger is a very useful building block for detecting two discrete voltage levels and finds many uses in circuit designs.

Voltage controlled oscillator

By using two CA3080s and some 741 op-amps it is possible to make an oscillator, the frequency of which is voltage controllable. This unit finds many applications in the fields of electronic music production and test equipment.

The circuit (Figure 8) has been given a logarithmic control law, that is, the frequency of operation doubles for every volt increase in the control voltage. This makes it ideal for musical applications where linear control voltages need to be converted into musical intervals (which are logarithmically spaced) and also for audio testing where frequencies are generally measured as logarithmic functions.

One CA3080, IC2, is an integrator. The I_{ABC} current that drives this IC is used to either charge or discharge C1. This produces triangular waveforms which are buffered by IC3, which then drives the Schmitt trigger IC4. The hysteresis levels for this device are fixed at +/−1.5 V, being determined by R6 and R7.

The output of the Schmitt trigger is fed back in such a way as to control the direction of motion of the integrator's output. If the Schmitt output is high, then the integrator will ramp upwards and vice versa.

Imagine that the integrator is ramping *upwards*. When the integrator's output reaches the *upper* hysteresis level, the Schmitt will flip into its *low* state, and the integrator will start to ramp downwards. When it reaches the *low* hysteresis level the Schmitt will flip back into its *high* state. Thus the integrator ramps up and down in between the two hysteresis levels.

The speed at which it does this, and hence the oscillating frequency, is determined by the value of I_{ABC} for IC2. The larger the current, the faster the capacitor is charged and discharged.

Two outputs are produced, a triangle wave (buffered) from IC3 and a squarewave (unbuffered) from IC4. If the squarewave output is loaded, then the oscillation frequency will change so a buffer is advisable. ▶

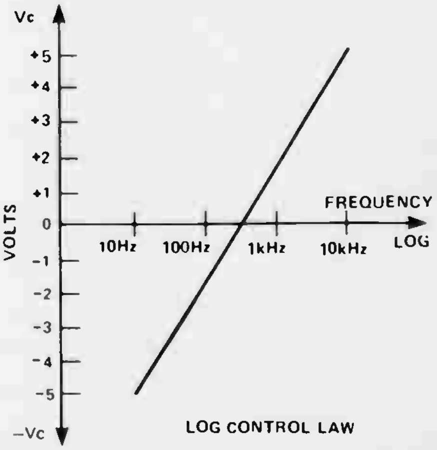


Figure 9. **Voltage versus frequency characteristic of the Figure 9 circuit.**

The log. law generator is composed of Q1, 2, 3 and IC1. Transistors Q1 and Q2 should be matched so that their base emitter voltages (Vbe) are the same for the same emitter current, (50uA). Matching these devices to within 5 mV is satisfactory, although unmatched pairs could be used. When matching transistors, take care not to touch them with your fingers. This will heat them up and produce erroneous measurements.

Transistor Q2 is used to produce a reference voltage of about −0.6 V, which is connected to IC1 pin 3. This op-amp and Q3 is used to keep the emitter of Q1 at the same voltage of −0.6 V. The input control voltage is attenuated by R1, R2 such that a +1 V increase at the input produces a change of only +18 mV at the base of Q1. However, the emitter of Q1 is fixed at −0.6 V, so the current through Q1 doubles. (It is a property of transistors that the collector current doubles for every 18 mV increase in Vbe).

The emitter current of Q1 flows through Q3 and into IC2, thus controlling the oscillator frequency. It is possible to get a control range of over 1000 to 1 using this circuit. With the values shown, operation from 10 Hz to 10 kHz is achieved. Reducing C1 to 1n will increase the maximum frequency to 100 kHz, although the waveform quality may be somewhat degraded.

Changing C1 to 1uF (non-polarised) will give a minimum frequency of 0.1 Hz.

Fast comparator

The high slew rate of the CA3080 makes it an excellent fast voltage comparator and a circuit is shown in Figure 10. When pin 2 of IC1 is more positive than Vref, the output of IC1 goes negative and vice versa. Vref can be moved around so that the point at which the output changes can be varied. As long as the input sinewave level is quite large (1 V say) then the output can be made to move at very fast rates indeed. However, care must be taken to avoid overloading the inputs. If the differential input voltage exceeds 5 V, then the input stage breaks down and may cause an undesired output to occur.

One use of a fast comparator is in a *tone burst generator*. A circuit is shown in Figure 11. This device produces bursts of sinewaves, the burst starting and finishing on axis crossings of the sinusoid. The CA3080 is configured here as a voltage comparator, used to detect these axis crossings and to produce a square wave output which then drives a binary divider (IC3). The divider produces a 'divide by sixteen' output which is high for eight sinewave cycles and then low for the next eight. This signal is then used to gate ON and OFF the sinewave.

The gate mechanism is a pair of transistors which short the sinewave to ground when the divider output is high and let it pass when the divider output is low. The resulting output is a toneburst.

However, if the comparator is not

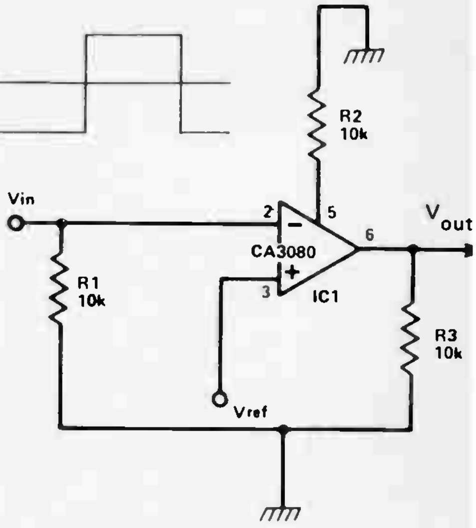


Figure 10. **Example of a fast comparator.**

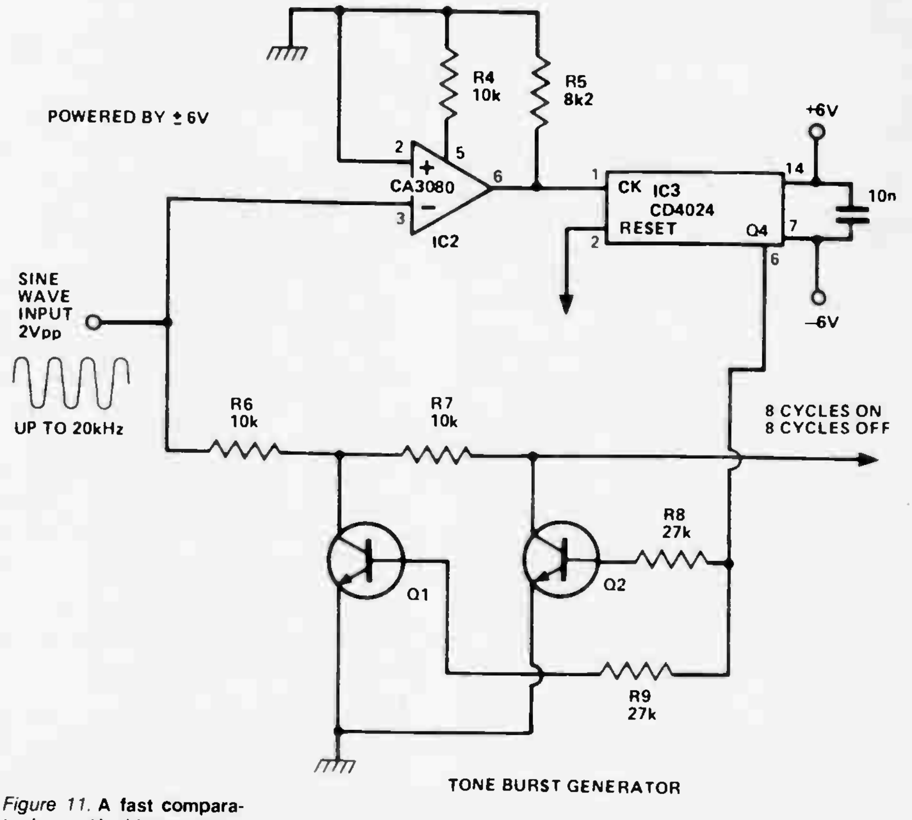


Figure 11. **A fast comparator is used in this tone burst generator, producing eight cycles of tone with eight cycle breaks starting and finishing at on-axis crossings.**

very fast then there will be a delay in generating the gate and so the tone burst will not start or finish on axis crossings.

Using the circuit shown, operation up to 20 kHz is obtainable.

Slew limiter

The current output of a CA3080 can be used to produce a controlled slew limiter. By connecting the output current to a capacitor, the output voltage cannot move faster than a rate given by

$$\text{Slew Rate} = \frac{I_{ABC}}{C1} \text{ Volts per sec.}$$

Note that I_{ABC} determines the slew rate and as I_{ABC} is a variable then so is the slew rate.

A suitable circuit is shown in Figure 12. The output voltage is buffered by a voltage follower, IC2. This is a MOSFET op-amp which has a very high input impedance, which is necessary to minimise the loading on C1.

When an input signal is applied to IC1 the output tries to move towards this voltage but its speed is limited by the slew rate. Thus, the output produces a linear ramp which stops when it reaches the input signal level.

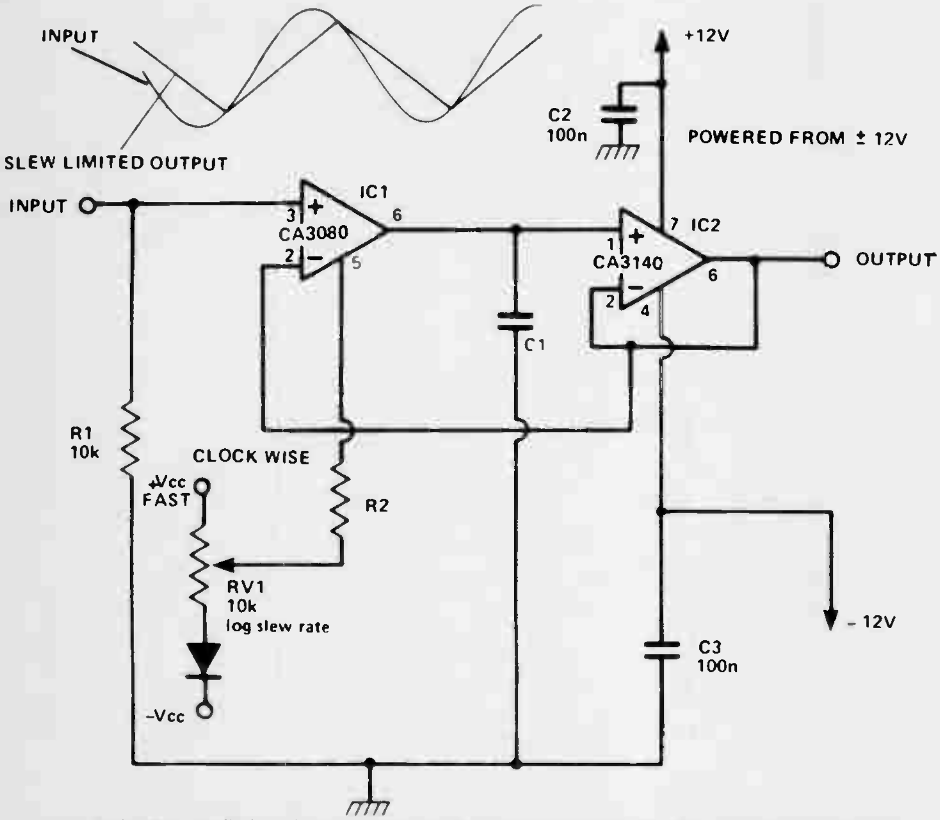


Figure 12. This slew rate limiter circuit produces a linear ramp on signals which exceed the slew rate limit, the output amplitude stopping when it reaches the signal level.

Sample and hold

A typical application of the slew limiter circuit is in a *sample and hold* circuit. The circuit in Figure 13 could be termed an analogue memory. When the control voltage is high, the circuit will 're-member' or 'hold' the input voltage level present at the time. The result is shown in Figure 14. ▶

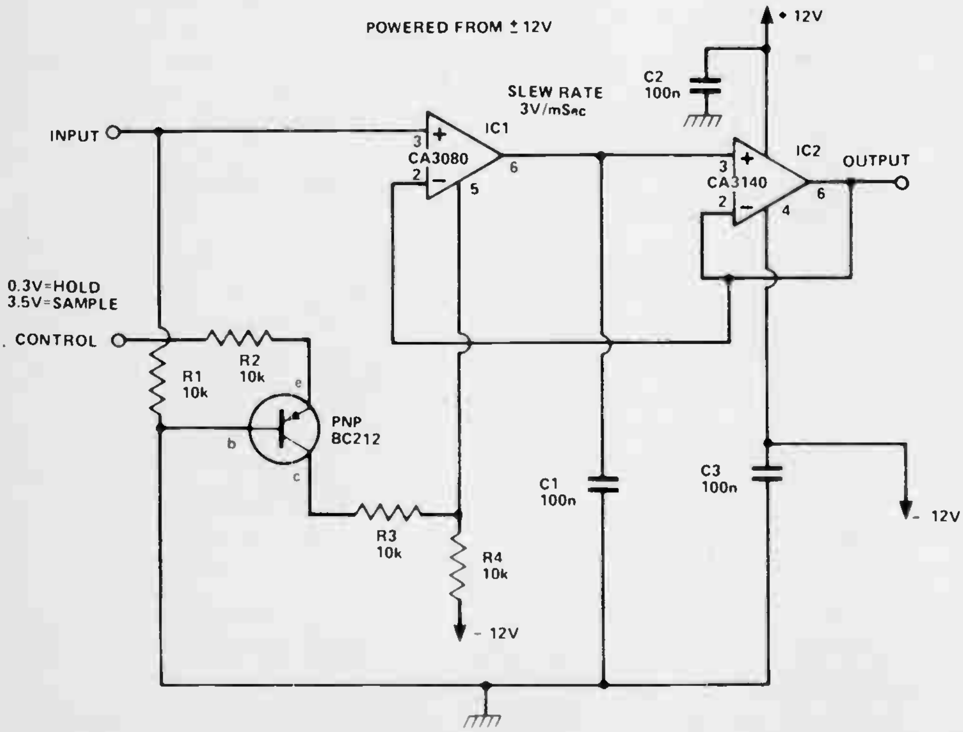


Figure 13. A typical application of the slew rate limiter is this sample and hold circuit.

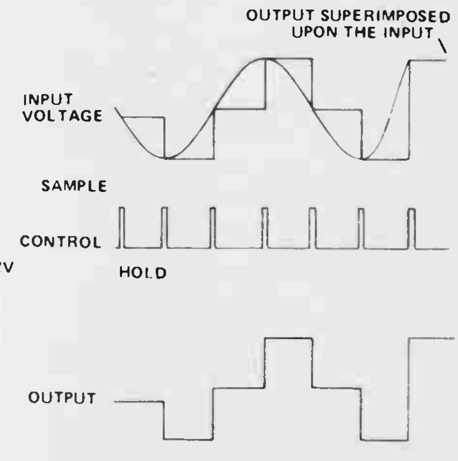


Figure 14. Illustrating the operation of the sample and hold circuit of Figure 13.

In this circuit, I_{ABC} is either hard ON (sample) or completely OFF (hold). In the sample mode, the output voltage quickly adjusts itself so that it equals the input voltage. This enables a short sample period to be used.

In the HOLD mode, I_{ABC} is zero and so the voltage on C1 should remain fixed.

Such circuits are used in music synthesizers (to remember the pitch), in analogue-to-digital converters and many other applications.

A multiplier/modulator

The CA3080 is basically a two-quadrant multiplier, that is, it has two inputs, one of which can accept bipolar signals (positive and negative going) — the inverting or the non-inverting input — the other can only accept a unipolar signal — the control input, pin 5.

Whilst a two-quadrant multiplier is very useful in a wide variety of applications, a four-quadrant multiplier has extra advantages. For example, apart from amplitude modulation, it can perform frequency doubling and ring modulation. See Figure 16. Now, a four-quadrant multiplier has two inputs, both of which can accept bipolar signals. An example of a four-quadrant multiplier is a frequency converter in a radio receiver. The familiar diode ring mixer is another example of a four-quadrant multiplier.

The circuit in Figure 15 is fairly similar to that of the two-quadrant multiplier shown in Figure 1. This circuit has several important differences.

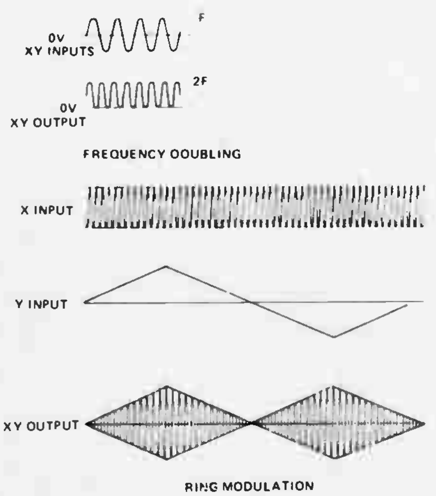


Figure 16. **Illustrating the various operations of the four quadrant multiplier of Figure 15.**

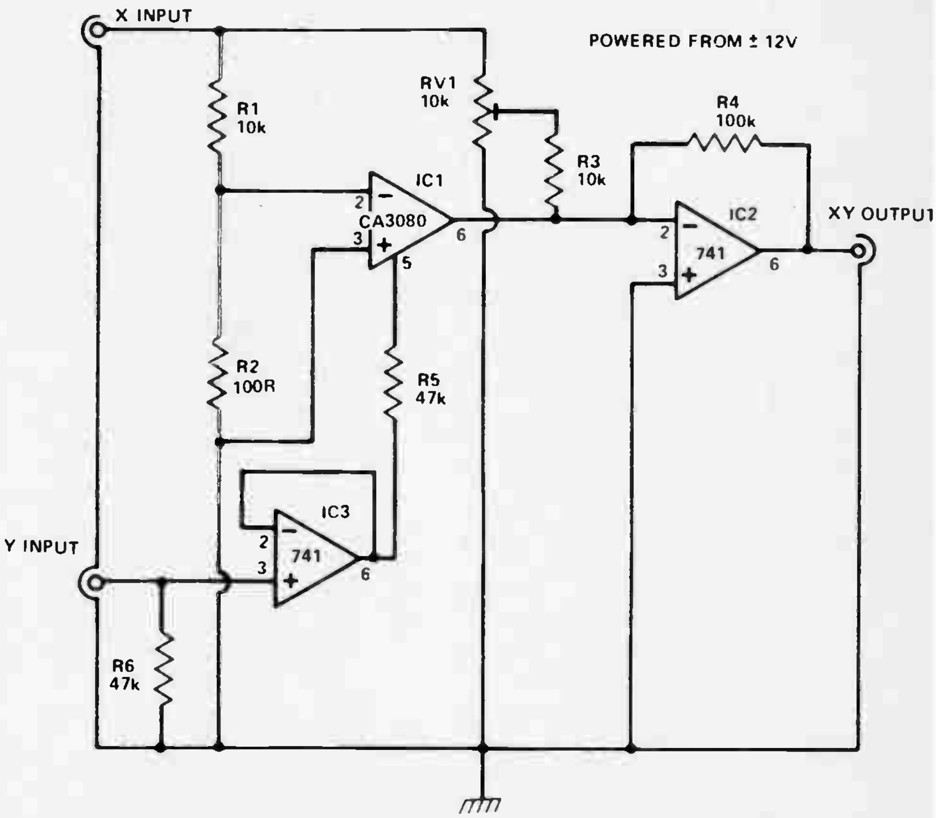


Figure 15. **This multiplier/modulator can be used to produce a 'Dalek' voice when working as a ring modulator. It can also be used as a frequency doubler.**

A 741 op-amp, IC3, is used to generate I_{ABC} in such a way that its input, the 'Y' input, can go both positive and negative. Thus, the Y input is bipolar.

When Y is at zero volts (no input) and there is a signal on the X input the desired output (X x Y) should be zero. This is achieved by adjusting RV1 so that the signal via IC1 (this is inverted) is exactly cancelled out by that via R3. Now, when Y is increased positively, a non-inverted value of X is produced at the output and, when Y is increased negatively, an inverted value of X is produced. When Y is zero, so is the output. This is known sometimes as ring modulation.

If a speech signal is connected to the X input and an audio oscillator to the Y input, the resulting sound is that of a 'Dalek'.

Also, if a sinewave is connected to both the X and Y inputs, the XY product is a sinewave of twice the frequency. This is known as a frequency doubler, but it will only work with sinewaves.

For more theoretical information on four-quadrant multipliers, especially the variable transconductance type, see "Operational Amplifiers" (second edition), by G.B. Clayton, published by Newnes-Butterworths

Single pole filter/wah wah

The guitar 'wah wah' effects unit employs a filter which can be manually 'swept' across the middle of the audio frequency range, generally from around 500 Hz to 5 kHz or so, producing the peculiar 'wah wah' sound.

A single pole, voltage-controlled, low pass filter can be constructed using a CA3080 as a current-controlled resistor. The circuit is showh in Figure 17.

A simple, low pass RC filter configuration is employed, the controllable 'R' is the CA3080 and the 'C' is C1. Varying I_{ABC} varies the amount of current drive to C1. This circuit configuration would normally be a slew limiter, except that the signal level to the input of the CA3080 is kept deliberately low (R1 and R2 form a 100:1 attenuator) so that the IC operates in its linear mode. This enables it to look like a variable resistor.

When this resistor is varied, the break frequency of the filter also varies.

By applying some positive feedback around the filter (R6, C2) it is possible to produce a peaky filter response. The peak actually increases with frequency, producing the wah wah effect.

The circuit as shown can be swept from about 400 Hz at the lower extreme to about 4 kHz at the upper extreme. See Figure 18. ▶

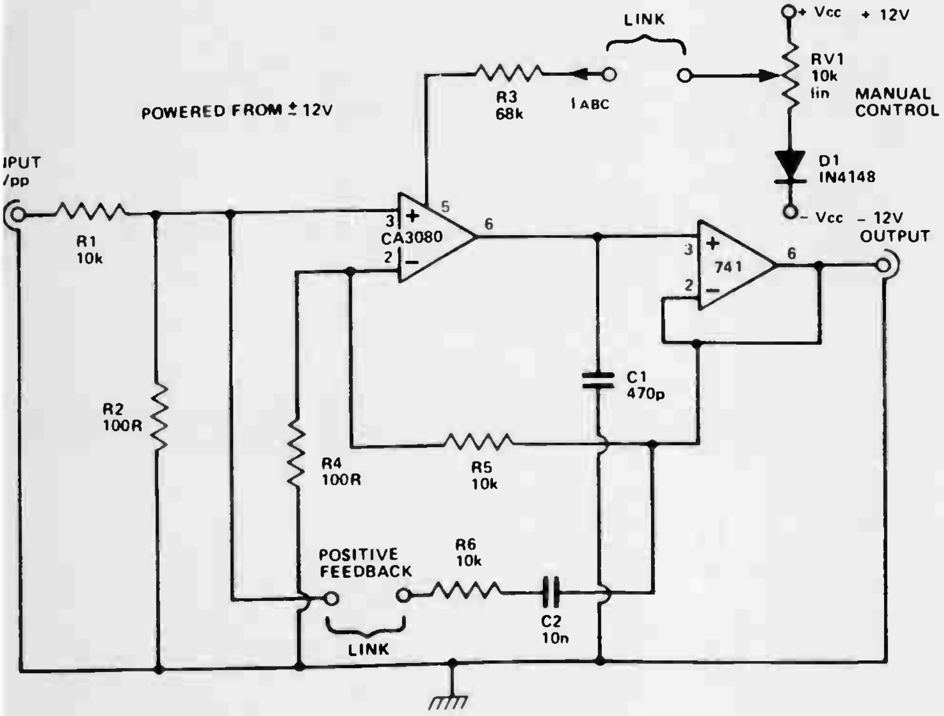


Figure 17. A guitar wah-wah unit can be made with a swept frequency single pole filter.

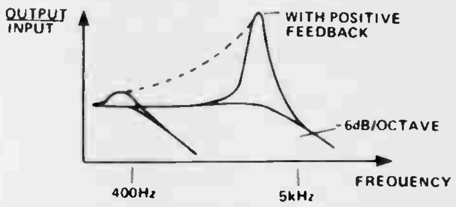


Figure 18. How the single pole filter affects the frequency response of the signal passed through the wah wah unit.

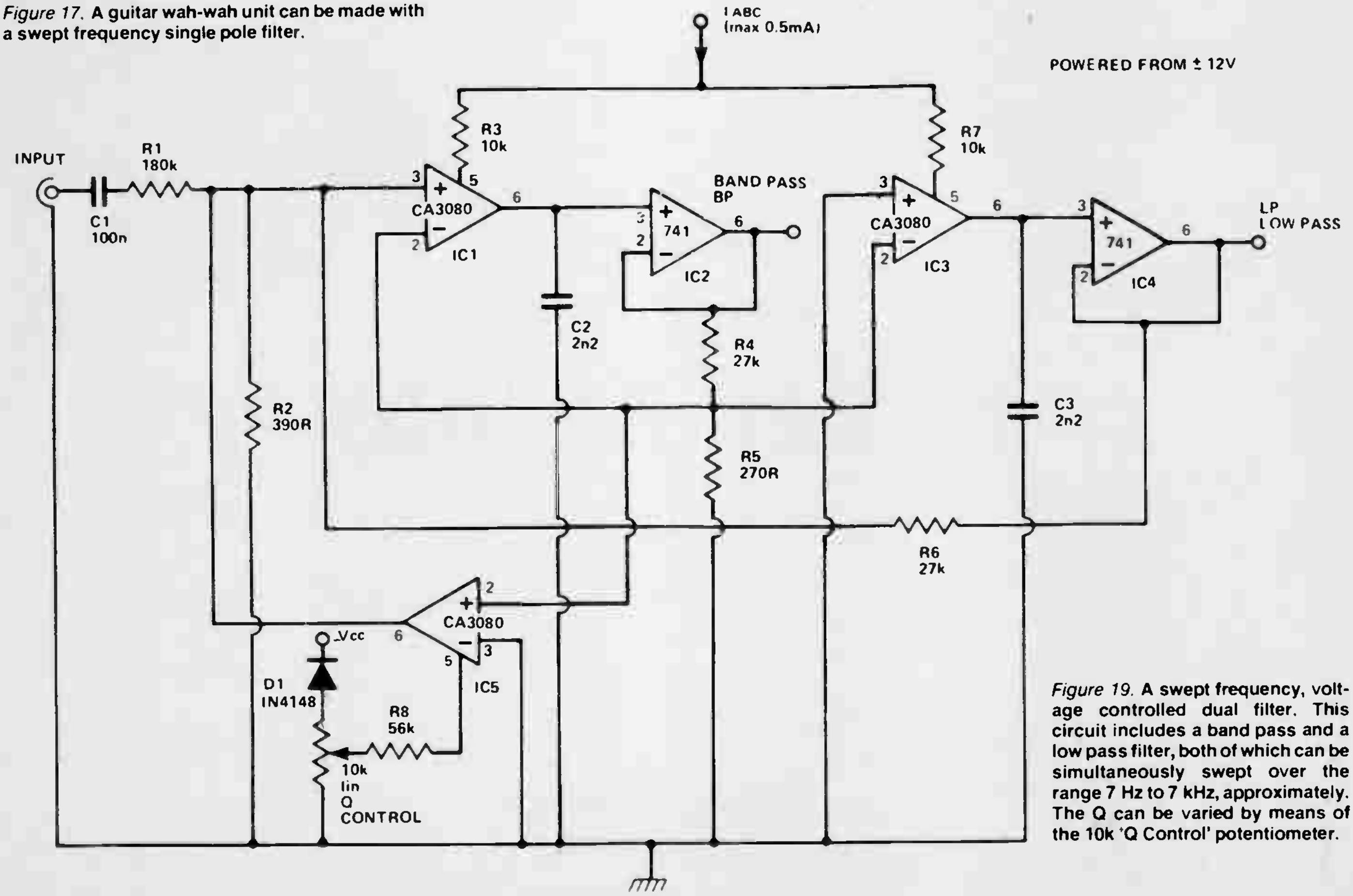
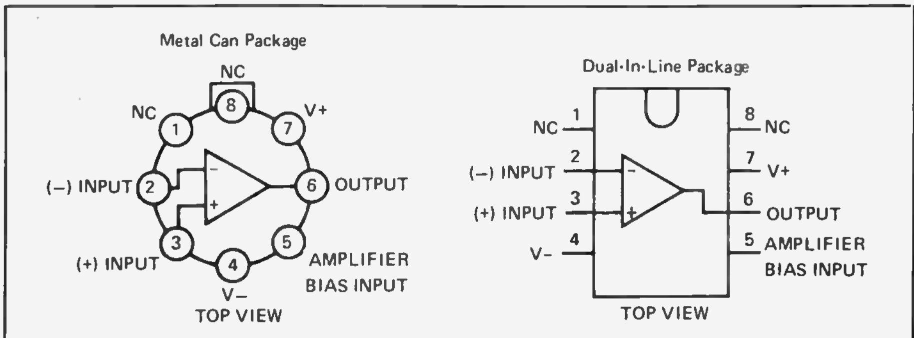


Figure 19. A swept frequency, voltage controlled dual filter. This circuit includes a band pass and a low pass filter, both of which can be simultaneously swept over the range 7 Hz to 7 kHz, approximately. The Q can be varied by means of the 10k 'Q Control' potentiometer.

Voltage controlled filter

A standard dual integrator filter can be constructed using a few CA3080s. By varying I_{ABC}, the resonant frequency can be swept over a 1000 to 1 range. IC1 and IC3 are two current-controlled integrators. IC2 and IC4 are voltage followers which serve to buffer the high impedance outputs of the integrators. A third CA3080 (C5) is used to control the Q factor of the filter. Q factors as high as 50 can be obtained. The resonant frequency of the filter is linearly proportional to I_{ABC} and hence this unit is very useful in electronic music production.

There are two outputs, a low pass and a band pass response. Minimum frequency is around 7 Hz to 10 Hz, upper frequency is around 7 kHz or so. Changing C2 and C3 will alter the upper and lower frequency limits. ●



PIN CONNECTIONS FOR THE 3080 IC

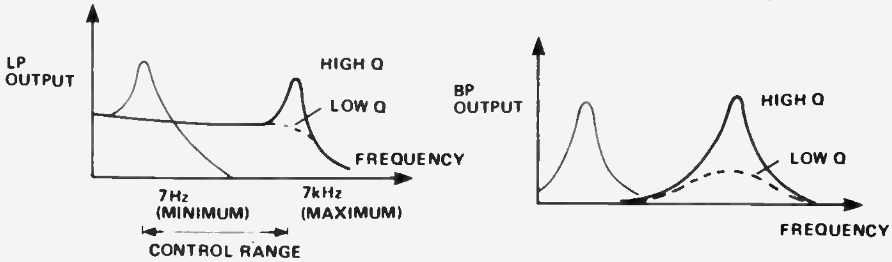


Figure 20. **Illustrating the operation of the filters in the Figure 19 circuit.**

Transistor arrays — using the 3046/3056/3086

Transistor arrays are a very useful electronic building block often overlooked by both hobbyists and professionals. This article describes a variety of circuits that can all be built using the 3046/3056/3086 — a common IC.

Peter Single

A transistor array consists of several transistors on the same piece of silicon. They are generally mounted in a 14- or 16-pin dual-in-line or T05 package. Because all the transistors are manufactured together, they are very well matched and because they are all on the same piece of silicon their thermal tracking is excellent.

Arrays may contain NPN transistors, 'super beta' NPNs, PNP transistors, zener diodes, transmission gates, SCRs and PUTs. The PNPs found in arrays are usually of inferior quality, having low β, low current handling capacity and poor frequency response. They are really only suitable for bias networks and similar low frequency applications.

The substrate connection

All transistor arrays have a connection labelled 'substrate'. Often it is connected to the emitter at an NPN transistor, or it may simply be brought out to a pin on the IC. A cross section of two transistors in an array is shown in Figure 1. The substrate consists of P-type silicon which isolates the two transistors shown as long as the collector-to-substrate junctions remain reverse biased. This is achieved by connecting the substrate to the most negative part of the circuit.

Beware of tricks which use the collector-to-substrate diode. They are dangerous! Figure 2 shows why. The substrate forms a PNP transistor with the base of the NPN transistor, which generates a latch. This latch can be active unless the PNP transistor is held off by taking the substrate connection to the most negative part of the circuit. Remember, when designing with arrays, BEWARE THE SUBSTRATE CONNECTION!

Another point to watch is that if the power supply is connected the wrong way round the collector-to-substrate diode may be forward biased, which can often destroy the array.

An internal schematic and pin connection diagram for the 3086 is shown on page 63. A similar diagram is shown in the RCA Data Book, but with one significant error! *The substrate connection is pin 13 not pin 12.*

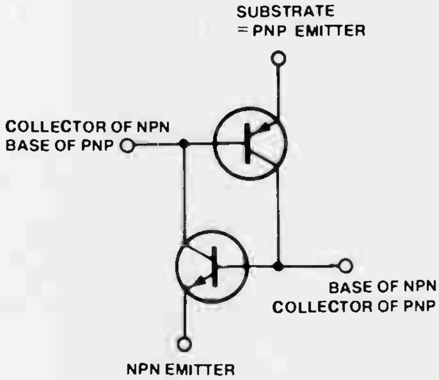


Figure 2. **Beware the collector-to-substrate diode!**

Current mirrors

A very useful circuit configuration that is easily implemented with arrays is the current mirror. Figure 3 shows the basic circuit.

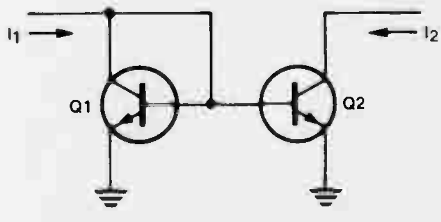


Figure 3. **Basic circuit of a current mirror. Connected like this, Q1 and Q2 on the array will have identical collector currents.**

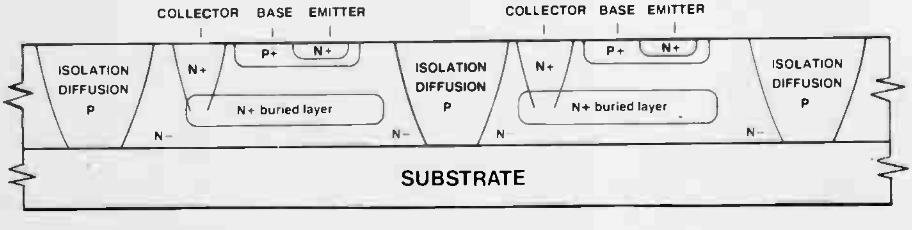


Figure 1. **Cross-section of two transistors in a transistor array IC.**

The collector current of a non-saturated transistor depends on the base-emitter voltage. If two identical transistors have identical base-emitter voltages their collector currents will be the same. In Figure 3, Q1 is diode connected and presents a low impedance to the input current. Transistor Q2, whose collector current mirrors that of Q1, is ▶

SELECTED DATA ON THE 3045/3046/3086 TRANSISTOR ARRAYS

General Description

The 3045, 3046 and 3086 each consist of five general purpose silicon NPN transistors on a common monolithic substrate. Two of the transistors are internally connected to form a differentially-connected pair. The transistors are well suited to a wide variety of applications in low power system in the dc through VHF range. They may be used as discrete transistors in conventional circuits however, in addition, they provide the very significant inherent integrated circuit advantages of close electrical and thermal matching. The 3045 is supplied in a 14-lead cavity dual-in-line package rated for operation over the full military temperature range. The 3046 and 3086 are electrically identical to the 3045 but are supplied in a 14-lead moulded dual-in-line package for applications requiring only a limited temperature range.

features

- Two matched pairs of transistors
  V_{BE} matched ±5 mV
  Input offset current 2μA max at I_C = 1 mA
- Five general purpose monolithic transistors
- Operation from DC to 120 MHz
- Wide operating current range
- Low noise figure 3.2 dB typ at 1 kHz
- Full military temperature range (3045) −55°C to +125°C

applications

- General use in all types of signal processing systems operating anywhere in the frequency range from DC to VHF
- Custom designed differential amplifiers
- Temperature compensated amplifiers

absolute maximum ratings (T_A = 25°C)

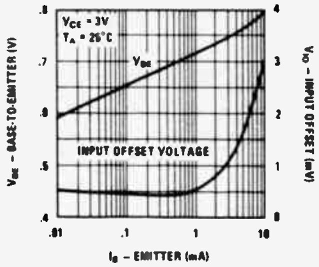
	3045 Each Transistor	3045 Total Package	3046/3086 Each Transistor	3046/3086 Total Package	Units
Power Dissipation:					
T_A = 25°C	300	750	300	750	mW
T_A = 25°C to 55°C			300	750	mW
T_A > 55°C			Derate at 6.67		mW/°C
T_A = 25°C to 75°C	300	750			mW
T_A > 75°C	Derate at 8				mW/°C
Collector to Emitter Voltage, V_{CEO}	15		15		V
Collector to Base Voltage, V_{CBO}	20		20		V
Collector to Substrate Voltage, V_{CIO} (Note 1)	20		20		V
Emitter to Base Voltage, V_{EBO}	5		5		V
Collector Current, I_C	50		50		mA

electrical characteristics (T_A = 25°C unless otherwise specified)

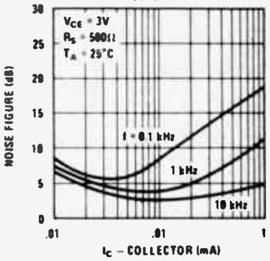
PARAMETER	CONDITIONS	LIMITS 3045, 3046 MIN	TYP	MAX	LIMITS 3086 MIN	TYP	MAX	UNITS
Static Forward Current Transfer Ratio (Static Beta) (h_{FE})	V_{CE} = 3V, I_C = 10 mA		100			100		
	V_{CE} = 3V, I_C = 1 mA	40	100		40	100		
	V_{CE} = 3V, I_C = 10 μA		54			54		
Input Offset Current for Matched Pair Q_1 and Q_2 $\lvert I_{O1} - I_{O2}\rvert$	V_{CE} = 3V, I_C = 1 mA		3	2				μA
Base to Emitter Voltage (V_{BE})	V_{CE} = 3V, I_E = 1 mA		715			715		V
	V_{CE} = 3V, I_E = 10 mA		800			800		V
Magnitude of Input Offset Voltage for Differential Pair $\lvert V_{BE1} - V_{BE2}\rvert$	V_{CE} = 3V, I_C = 1 mA		45	5				mV
Magnitude of Input Offset Voltage for Isolated Transistors $\lvert V_{BE3} - V_{BE4}\rvert$, $\lvert V_{BE4} - V_{BE5}\rvert$, $\lvert V_{BE5} - V_{BE3}\rvert$	V_{CE} = 3V, I_C = 1 mA		45	5				mV
Temperature Coefficient of Base to Emitter Voltage $\left(\frac{\Delta V_{BE}}{\Delta T}\right)$	V_{CE} = 3V, I_C = 1 mA		−1.9			−1 9		mV/°C
Collector to Emitter Saturation Voltage ($V_{CE(SAT)}$)	I_B = 1 mA, I_C = 10 mA		23			23		V
Temperature Coefficient of Input Offset Voltage $\left(\frac{\Delta V_{IO}}{\Delta T}\right)$	V_{CE} = 3V, I_C = 1 mA		1 1					μV/°C

Note 1: The collector of each transistor of the 3045, 3046, and 3086 is isolated from the substrate by an integral diode. The substrate (terminal 13) must be connected to the most negative point in the external circuit to maintain isolation between transistors and to provide for normal transistor action.

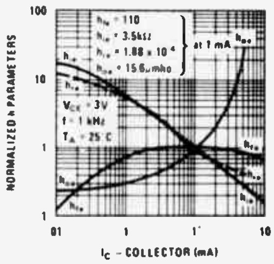
Typical Static Base To Emitter Voltage Characteristic and Input Offset Voltage for Differential Pair and Paired Isolated Transistors vs Emitter Current



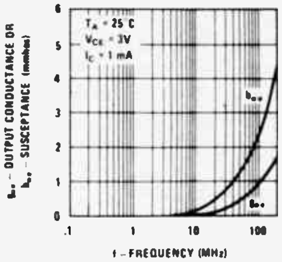
Typical Noise Figure vs Collector Current



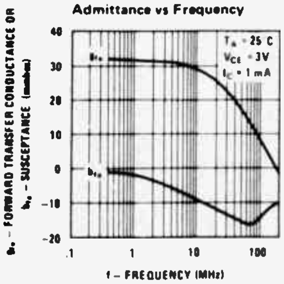
Typical Normalized Forward Current Transfer Ratio, Short Circuit Input Impedance, Open Circuit Output Impedance, and Open Circuit Reverse Voltage Transfer Ratio vs Collector Current



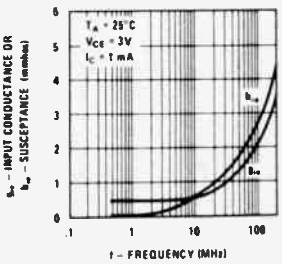
Typical Output Admittance vs Frequency



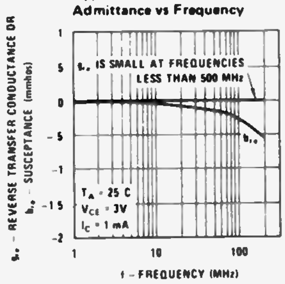
Typical Forward Transfer Admittance vs Frequency



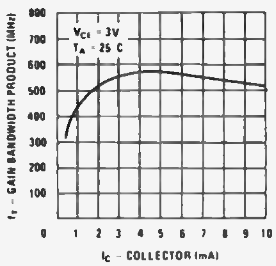
Typical Input Admittance vs Frequency



Typical Reverse Transfer Admittance vs Frequency



Typical Gain-Bandwidth Product vs Collector Current



connected as a dc-coupled common emitter amplifier and has a very high output impedance.

Current mirrors constructed from discrete components are not nearly as accurate because two discrete transistors chosen at random will rarely be identical. Also it requires extra construction to ensure that the transistors are held at the same temperature, because transistor characteristics are very temperature-dependent.

But two transistors on an array, having been manufactured side by side and being incorporated in the same piece of silicon, are very well matched. Errors in mirrors constructed using arrays are generally less than ten per cent.

A current mirror provides a regulated high impedance current sink. Its output will generally go to 200 mV from earth before the mirror transistor saturates, making it very useful for linear ramps, biasing networks and a host of other applications.

Voltage controlled oscillator

A circuit for a voltage controlled oscillator is shown in Figure 4. It consists of a controlled bias system with Q2 and Q3 forming a current mirror with Q1. To understand the oscillator, imagine that Q5 is turned hard on. Transistor Q2 will slowly pull Q4's emitter low until Q4 starts to turn on. As it does so it will pull Q5's base low, turning Q5 off which turns Q4 harder on. This regenerative action continues until Q4 is fully on and Q5 off. Q3 then slowly starts to turn Q5 on and the process continues.

The rate of oscillation depends on the charging time of the capacitor between the collectors of Q2 and Q3. If the current through Q2 and Q3 decreases, the charging time increases and the oscillator frequency decreases. Because Q2 and Q3 mirror Q1, varying the current through Q1 will vary the oscillator frequency.

The circuit will oscillate at about 10 MHz with 5 V on the control pin. It can be made to run at 30 MHz by placing germanium diodes between collector and base of Q4 and Q5 and increasing the control voltage. The diodes should be inserted with their cathodes to the transistor bases to stop the transistors from saturating and storing charge in their bases. Such charge storage slows transistors down.

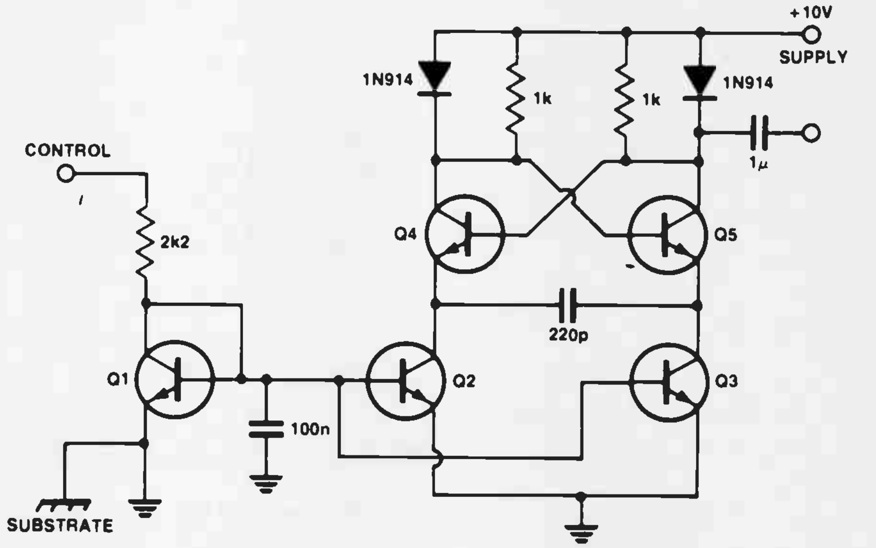


Figure 4. **Circuit of a voltage controlled oscillator. Transistors Q2 and Q3 in the array form a current mirror with Q1. The voltage at the 'control' input sets the collector currents of Q2 and Q3 which controls the oscillator frequency. Circuit oscillates around 10 MHz with 5 V on the control input.**

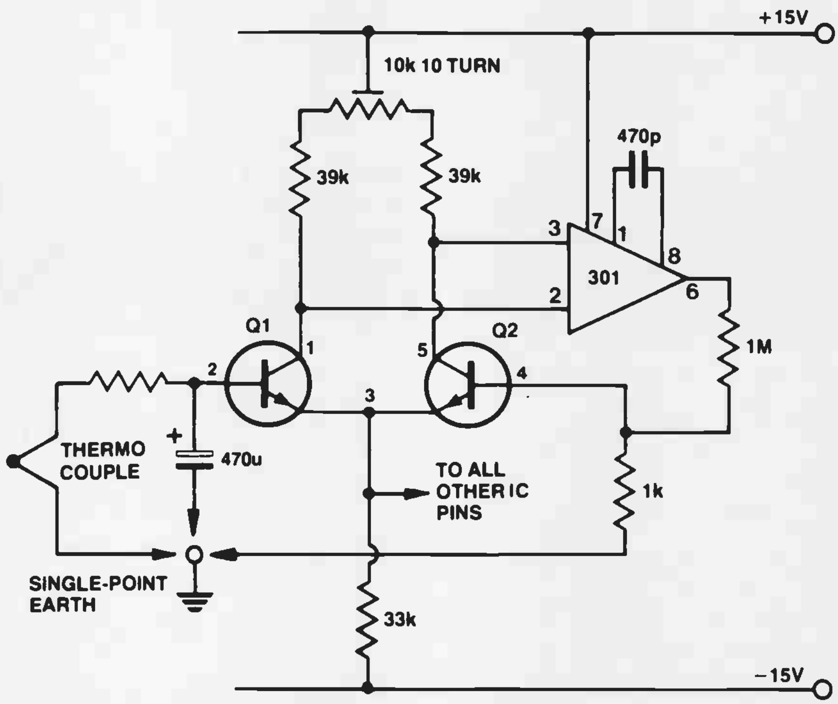


Figure 5. **A useful thermocouple amplifier. Input offset nulling is independant of temperature in this circuit, unlike many circuits.**

Thermocouple amplifier

Figure 5 shows a circuit for an amplifier with a gain of 1000, suitable for use with thermocouples.

The biggest problem when using op-amps as thermocouple amplifiers is the temperature drift of the input offset voltage. Most methods of nulling the input offset are ineffective if the temperature varies. This circuit solves the problem by eliminating the input offset voltage ▶

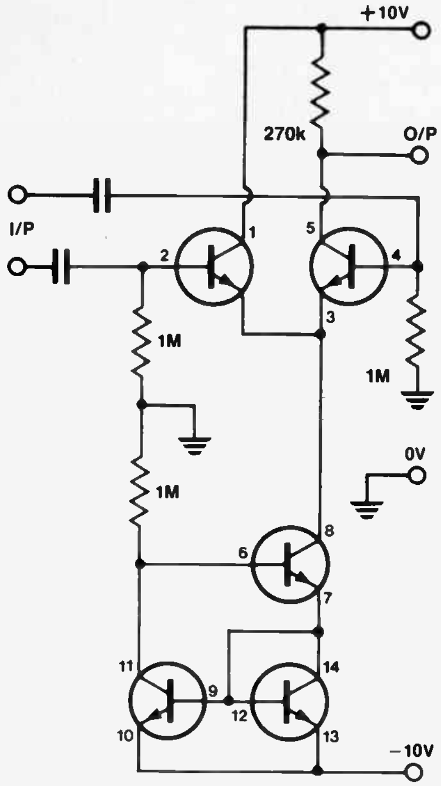


Figure 6. **This differential amplifier has a very high common mode rejection ratio (cmrr). Constant current source for the differential pair is a variation on the current mirror. Circuit is suitable for bio-electronic applications such as in a cardiac monitor or EMG.**

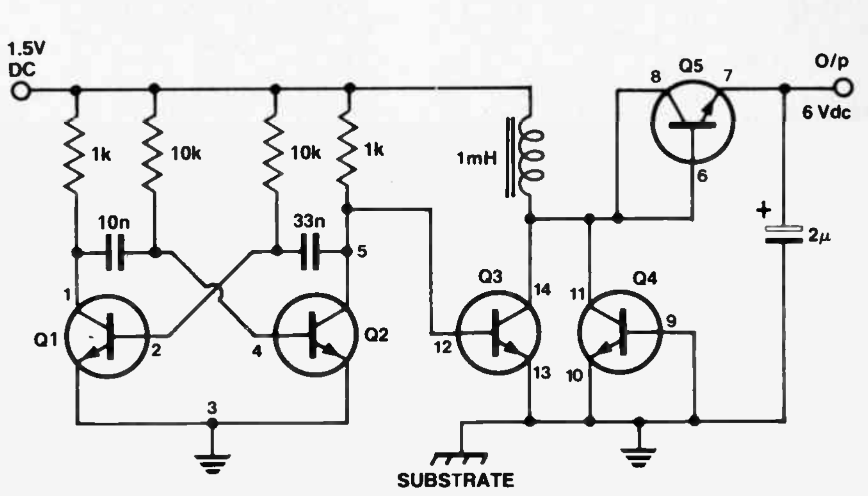


Figure 7. **This ringing choke inverter is just the thing for powering CMOS circuits from a single 1.5 V cell in portable applications. Operation is explained in the text.**

entirely. With a piece of wire replacing the thermocouple adjust the pot until the output of the amplifier is at 0 V. Since the bases of both Q1 and Q2 are at earth there is no input offset voltage.

This amplifier has very high open loop gain and will happily amplify any hum or other noise at its input. It will also oscillate through any capacitive coupling between output and input or between the supplies and the input. To avoid this, supplies should be well bypassed and earthing should be 'star' configured.

The substrate is not connected to the negative supply, because under extreme conditions the 20 V maximum collector-to-substrate voltage could be exceeded. However, no part of the array will ever go more negative than the substrate, so the circuit is quite safe.

The other transistors on the array should not be used as they may generate temperature gradients between Q1 and Q2.

Differential amplifier

The circuit shown in Figure 6 has a very high common mode rejection ratio and is suitable for circuits such as cardiac monitors where the desired signal may be buried in a lot of common mode noise. It's a differential amplifier biased by a variation of a current mirror that has very high output impedance to improve the common mode rejection.

Ringing choke inverter

This is a simple circuit, illustrated in figure 7, which generates about 6 V at 50 mA from a 1.5 V battery. It could be used to power CMOS from a single penlight battery when a compact, portable circuit is required.

Transistors Q1 and Q2 form an astable multivibrator. When Q2 turns off, Q3 turns on and current flows through the inductor from the supply. The point 'A' on the waveform diagram (Figure 8) is the point where Q3 saturates. When Q2 turns on, cutting off Q3, the inductor tries to maintain the flow of current. Q5 is connected as a diode and current will flow through it until the voltage on Q4's emitter zeners the reverse biased junction. This is point 'B' on Figure 8.

When the inductor no longer has the energy to pump current into Q4 or Q5 it starts to oscillate (or ring) with its own parallel parasitic capacitance. This is point 'C' on Figure 8. The amplitude of

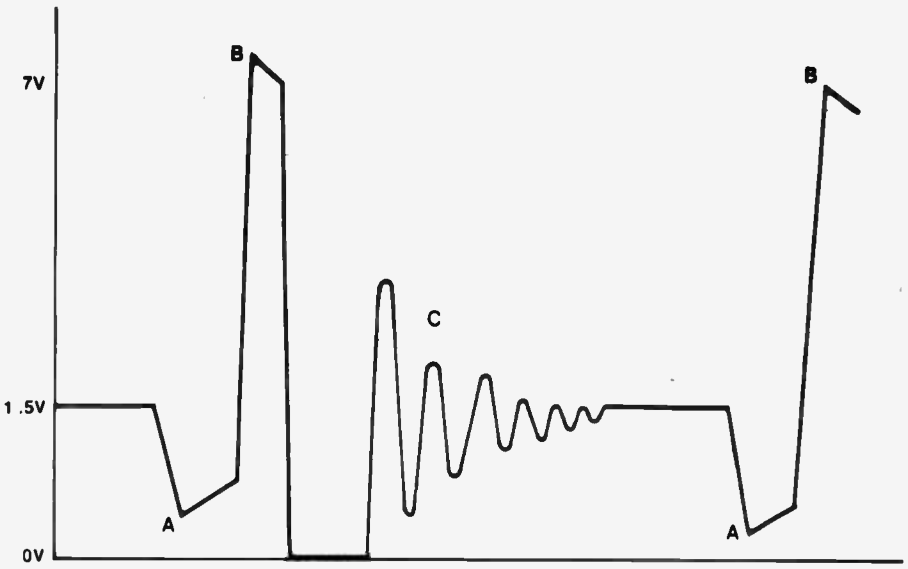


Figure 8. **Typical waveform at the collector of Q3 in the ringing choke inverter.**

the selection of the correct heatsink, the procedure is as follows.

Calculate the *worst case continuous average power dissipation* in the regulator from the formula:

$P = (V_{in} - V_{out}) \times I_{out}$.

The voltage/current characteristics of the unregulated input must be accurate. A small change in input voltage can result in a large increase in the power dissipated by the regulator. For example, normal operating conditions are:

$V_{out} = 10$ V
$V_{in} = 14$ V
$I_{out} = 10$ A
$P = (14 - 10) \times 10$
$= 40$ watts.

If the input voltage increases by 10% to 15.4 volts:

$P = (15.4 - 10) \times 10$
$= 54$ watts

— an increase in power dissipation of 35%.

Therefore, the power supply circuit up to the regulator input (i.e: transformer, rectifier diodes, filter capacitor) plays an important role in the successful operation of the regulator itself. It should be built and tested to determine its average dc output voltage under full load with maximum input voltage. This circuit is shown in Figure 2.

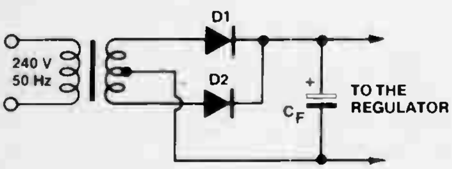


Figure 2. **Circuit prior to the regulator.**

The choice of C_F is also very important. At *high* current levels the capacitor ripple current (RMS) is two to three times the dc output current. If the capacitor has an equivalent series resistance (ESR) of 0.05 ohms, this can cause internal power dissipation (I^2R) of 20 to 45 watts at an output current of 10 amperes.

The life of the capacitor 'derates' with increase in operating temperature, and the choice of a small-value capacitor is asking for trouble (about 2000 μF is used for the LM 317 circuit). A value of some *2000 μF per ampere of load current* is the minimum recommended value. Large values of capacitor will have longer life and will also reduce the ripple level. This allows a lower dc input voltage to the regulator, which will result in savings in transformer and heatsink costs.

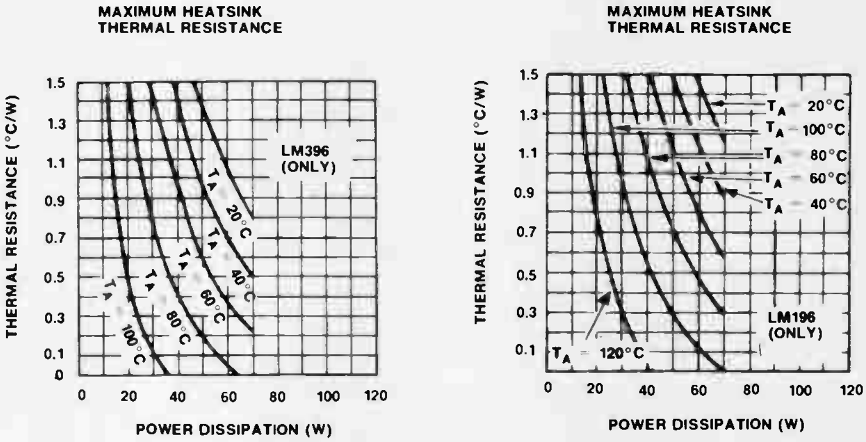


Figure 3. **Heatsink thermal resistance graphs (T_A = Ambient temperature)**

A further idea is to place several capacitors in parallel. This increases the capacitance, reduces the net series resistance and increases the heat dissipating area (i.e: shares it among the capacitors). Once the circuit in Figure 2 has been finalised and the average dc output voltage determined, the thermal resistance of the heatsink can be determined from the graphs in Figure 3, in degrees centigrade per watt (°C/W).

For conservative heatsinking it is recommended that you choose T_A to be 35°C higher than anticipated.

The heatsink resistance generally falls into the range of 0.2°C/W - 1.5°C/W at a T_A = 60°C. These are *large* heatsinks such as the Philips 45D6CB, 55D6CB, and the large Minifin. These must be mounted for best convection cooling and could also be cooled by a fan.

2. Transformers

Correct transformer ratings are extremely important in high current supplies. If the secondary voltage is too high, power will be wasted and cause unnecessary power dissipation in the regulator. However, if the secondary voltage is too low it may cause loss of regulation if the input voltage (i.e: mains) fluctuates excessively.

The following formula can be used to calculate the secondary voltage required using the circuit in Figure 2 (full wave centre tap).

$$V_{(RMS)} = \frac{V_{out} + V_{reg} + V_{Rect} + V_{Ripple}}{\sqrt{2}} \times \frac{V_{Nom}}{V_{Low}} \times (1.1) \qquad (1)$$

Where:

1.1 is the factor accounting for load regulation of the transformer.
V_{out} = dc regulated output voltage.
V_{Reg} = Minimum $V_{in} - V_{out}$.
V_{Rect} = Voltage drop (forward) across the diode at $3 \times I_{out}$
V_{Ripple} = Peak capacitor ripple voltage (½ p-p).

i.e: $\dfrac{(5.3 \times 10^{-3})\ I_{out}}{2C}$

C is the capacitor value in farads.
V_{Nom} = Normal ac input (RMS).
V_{Low} = Minimum ac input (RMS).

The current rating required can be calculated from the formula:

$$I_{RMS} = I_{out} \times 1.2 \qquad (2)$$

Where I_{out} = dc output current.

Transposing formula (2) we can calculate the value of filter capacitor required:

$$C = \frac{(5.3 \times 10^{-3})\ I_{out}}{2 \times V_{Ripple}} \qquad (3)$$

The best way to appreciate these formulas in use is to calculate the values required for a power supply circuit. If we design a good mobile radio power supply, 13.8 volts at 10 amperes: ▶

V_{out} = 13.8 V
I_{out} = 10 A
Assume V_{Reg} = 2.2 V, V_{Rect} = 1.2 V
V_{Ripple} = 2 V p-p, V_{Nom} = 240 V
V_{Low} = 220 V

Using formula (1)

$$V_{(RMS)} = \left(\frac{13.8 + 2.2 + 1.2 + 1}{\sqrt{2}}\right)\left(\frac{240}{220}\right) 1.1$$

$$= \frac{18.2}{\sqrt{2}} \times 1.09 \times 1.1$$

$$= 12.869 \times 1.09 \times 1.1$$

$$= \underline{15.4 \text{ volts (RMS)}}$$

Using formula (2)

$$I_{(RMS)} = 10 \times 1.2$$

$$= 12 \text{ amperes (RMS)}$$

The transformer must therefore be 240: 30 CT at 12 amperes. The centre tap (CT) will provide 15 volts secondary voltage for each diode.

The size of the filter capacitor required can be calculated using formula (3)

$$C = \frac{(5.3 \times 10^{-3})\ 10}{2 \times 1}$$

$$= 26500\ \mu F$$

The transformer, rectifier and filter circuit is now shown in Figure 4.

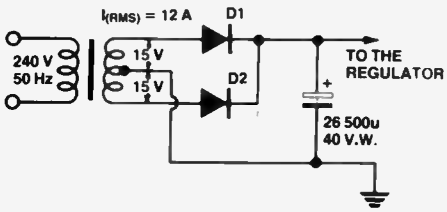


Figure 4. **Rectifier and filter circuit**

3. Diodes

The diodes used in the circuit must have a high dc current rating. The capacitor input filter draws high peak current pulses that are considerably higher than the average dc current. With a 10 amperes supply the average current is 5 amperes. The current pulses' duration and amplitude result in a long-term diode heating of approximately 10 amperes dc. Therefore the diodes should have a rating of at least 10-15 amperes. Also, the power supply may have to survive a short circuit and average current could rise to 15 amperes (see Figure 1a).

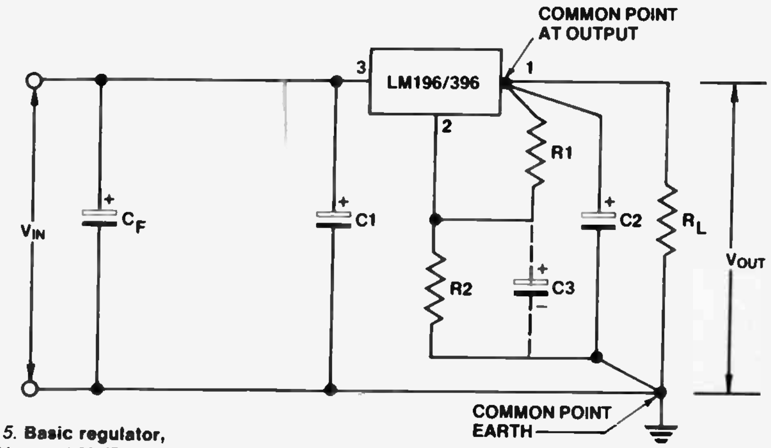


Figure 5. **Basic regulator,** $V_{out} = 1.25\ \frac{(R1 + R2)}{R1}$

Another important factor in the choice of diode is the surge current at switch on. The peak *surge current* is about *10-20 times the dc output current* (i.e: 100-200 A for a 10 A supply). (Note: smaller transformers and filter capacitors may be used in lower current supplies. This will reduce the surge current; unless you are sure of the worst case surges, do not economise on diodes.)

Stud-mounted diodes in a DO-4 or DO-5 package are recommended, such as IR 12F10B, IN3209 or 16F10 silicon rectifiers. Remember to choose the correct PIV for the type of transformer in use (PIV = $\sqrt{2}\ V_{Secondary}$).

4. Wiring

High load currents produce higher than normal voltage drops across the resistance of the wiring. It is suggested that 16-18 gauge wire is used for input and output connections, and the *length is kept to a minimum.*

The two resistors used to set the output voltage level are connected:

1. directly to a common point earth and
2. directly to the output of the regulator as shown in Figure 5.

Components in Figure 5.

C_F = Main filter capacitor 26 500 μF.

C1 = 4 μ7 tantalum. It is only necessary if the main filter capacitor is more than 150 mm away from the regulator. Connecting wire is 18 gauge or larger.

C2 = 4 μ7 tantalum. It is not absolutely necessary, but is recommended to maintain low output impedance at high frequencies.

C3 = 25 μF. Improves ripple rejection, output impedance, and noise. (Capacitor C2 should be close to the regulator if C3 is used).

R1 = 120 ohms. It should be a wirewound or metal film resistor, tolerance 1% or better.

R2 = calculated to set V_{out}; the same type of resistor as R1.

The value of R2 can be calculated from the formula:

$$R_2 = \left(\frac{V_{out}}{1.25}\right) \times R1 - R1$$

Example:

V_{out} = 13.8 V
R1 = 120 ohms

$$R2 = \left(\frac{13.8}{1.25}\right) \times 120 - 120$$

$$= (11.04 \times 120) - 120$$

$$= 1324.8 - 120$$

$$= 1204.8 \text{ ohms.}$$

As stated earlier, the package is a TO-3 and the connections are shown in Figure 6.

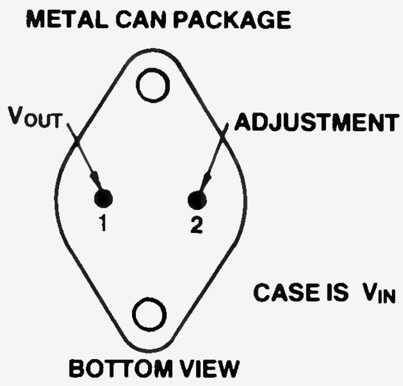


Figure 6. **Connection diagram**

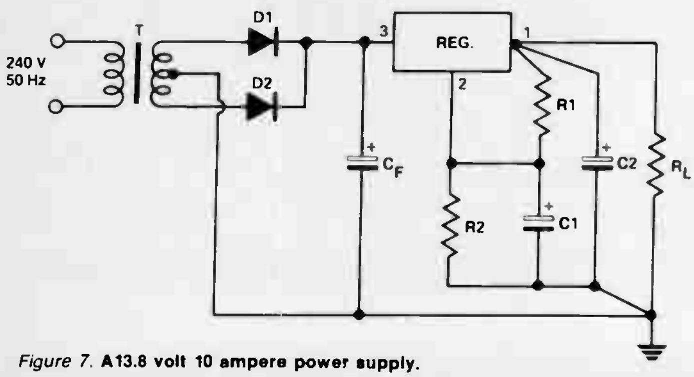


Figure 7. **A13.8 volt 10 ampere power supply.**

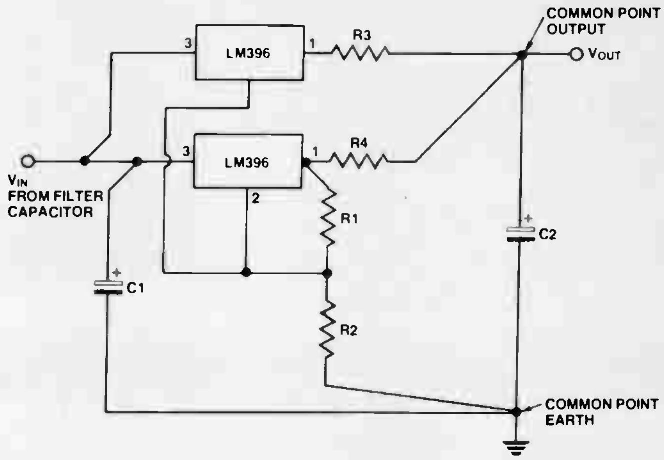


Figure 9. **Quasi-parallel regulators**
R1 = 120 ohms
R2 chosen to set V_{out}
R3, R4 = 0.015 ohms
C1 = 4μ7 tantalum
C2 = 100μF

The complete circuit can now be built, incorporating Figures 4 and 5. The circuit diagram of the final 13.8 V 10 A power supply is shown in Figure 7.

Component values for Figure 7.

T	=	240 : 30 CT at 12 amperes.
D1	=	16F10 DO-4 case.
D2	=	16F10 DO-4 case.
CF	=	26500 μF 40 VW (ideally, capacitors in parallel).
C1	=	25 μF 16 VW.
C2	=	4μ7 tantalum 16 VW.
R1	=	120 ohms 1% metal film.
R2	=	1k2 1% metal film.
Reg	=	LM396 on a 6" 55 or 65D heatsink.

$$V_{out} = 1.25\left(\frac{R1 + R2}{R1}\right) = 1.25\left(\frac{120 + 1200}{120}\right) = 1.25 \times 11 = 13.75 \text{ volts}$$

A highly desirable situation would be to *reduce* the power dissipated by the regulator. This can be achieved by supplying part of the output current around the regulator as shown in Figure 8.

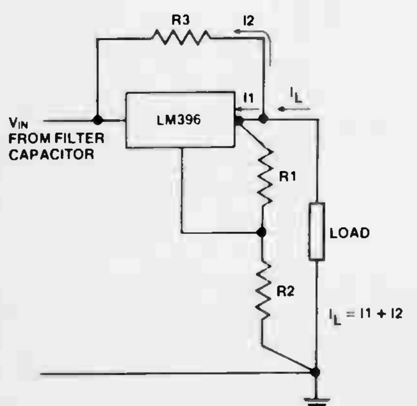


Figure 8. **Reducing regulator power dissipation**

Resistor R3 is selected to supply a portion of the load current. In this case a *minimum load must always be maintained.* This prevents the regulated output from rising uncontrolled. The value of R3 must be greater than:

$$\frac{V_{max} - V_{out}}{I\ min} \text{ ohms}$$

Where: V_{max} is worst case high input voltage.
I_{min} is the minimum load current.
Power rating must also be considered and R3 must be rated at a minimum of:

$$\frac{(V_{in} - V_{out})^2}{R3} \text{ watts}$$

This circuit configuration will reduce the regulator power dissipation by a factor of 2 to 3, if the minimum load current is about 50% of the full load current.

Precautions when using R3

1. The power rating of R3 must be *increased* to $\frac{(V_{max})^2}{R3}$ watts if continuous output short circuits are at all likely.
2. Under short circuit conditions the *overall circuit* power dissipation increases by $\frac{(V_{in})^2}{R3}$ watts.

The regulator and R3 will not be harmed (if R3 is the correct wattage), but the circuit components prior to the regulator (diodes, transformer) must be able to withstand the overload condition (i.e: the power rating is sufficient to handle the excess current).

The only problem with this technique is the large power rating required for resistor R3. If $V_{in} - V_{out} = 7$ volts and R3 = 2 ohms, the power dissipated by the resistor is:

$$\frac{(7)^2}{2} = 24.5 \text{ watts}$$

with 3.5 A of current passing through it.

High Current Output

Placing regulators in parallel is not recommended because they may not share the current equally. The regulator with the highest reference voltage will handle the highest current up to the time it current limits. Therefore, one regulator may be flat out handling 16 A while the other is cool and calm passing only 2 A. Reliability cannot be guaranteed under these conditions because of the high junction temperature of regulator one.

However, if load regulation is not critical, the regulators may be connected quasi-parallel, as shown in Figure 9. This circuit will share current to within 1 ampere, and in the worst case 3 amperes. However, the payoff is in the load regulation. It is degraded by 150 mV at 20 ampere loads compared to about 20 mV with 10 ampere loads. This should not cause too much of a problem in higher voltage power supplies. ●

Acknowledgement

This article was made possible by the courtesy of National Semiconductor. Data and basic circuits were taken from their publication 'LM196/LM396 10 Amp Moose Adjustable Voltage Regulator'.

Using BiFET and BiMOS op-amps

The availability of BIFET and BIMOS op-amps has revolutionised circuit design since they appeared on the scene five or so years ago. Here are practical applications circuits for a number of these devices.

Brian Dance

THE AVAILABILITY of BiFET and BiMOS devices in various packages with one to four operational amplifiers per package has revolutionised the operational amplifier market. Apart from the relatively expensive hybrid FET input devices, other FET input operational amplifiers had been available for some considerable time, so why should BiFET and BiMOS devices be so important?

The first point to note is that amplifiers with FET input stages can offer far higher input impedances than devices with ordinary bipolar transistors in their input stages. For example, the well-known 741 has an input impedance of the order of 1M and a maximum input bias current of 500 nA. The use of bipolar transistors to obtain a high input impedance has been pushed to the limit in devices such as the LM108, using supergain input devices to provide a typical input impedance of 70M and an input bias current of just under 1 nA. These values may be compared with those of some of the economical BiFET and BiMOS devices, where typical input impedances are of the order of 1 Terraohm (one million Megohms!) and input currents are some tens of picoamps (pA) at room temperature.

Thus if one connects the input of one of these BiFET or BiMOS amplifiers to almost any circuit, it will impose a very small load on that circuit. This can be a vital consideration when one is designing such high-impedance circuits as those used in pH meters or in ionisation chamber smoke detector circuits, whose output current is inadequate to drive devices such as the 741.

If one considers the very early types of monolithic FET input operational amplifiers (such as the Fairchild μA740), they do have the desired high input impedance, but their disadvantage is that their input offset voltage and its temperature coefficient are so high that they do not approach the high standard of performance required by the modern

INTRODUCTION TO THE BIMOS AND BIFET OP-AMP

The first BiFET products were announced by National Semiconductor in 1975 (the LF155, LF156 and LF157 series, where LF signifies Linear FET device). The main advantages of these products is that the junction FET devices used in their input stages are fabricated on the same silicon chip as the remainder of the operational amplifier. Although hybrid operational amplifiers with FET input stages had been available for some considerable time previously, all of these hybrid devices contained the junction FET devices fabricated on a separate silicon chip from the remainder of the operational amplifier. Such hybrid devices can be made to have a very good performance if adequate trouble is taken in their design, but the extra labour costs involved in the testing of the separate chips for appropriate matching characteristics and in connecting the two chips in a single hybrid package inevitably resulted in a price tag far above that of modern BiFET devices. The general type of construction of a BiFET device is shown in Figure 1, the channel between the source and the drain electrodes of the FET input devices being fabricated by ion implantation.

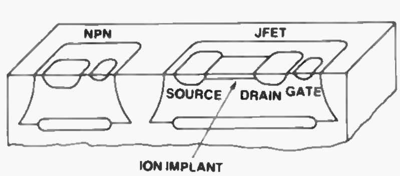


Figure 1. **Construction of a BiFET device.**

Although National Semiconductor produced the first BiFET products, it was not long before other manufacturers entered the BiFET market, and such products are now available from Advanced Micro Devices, Analog Devices, Fairchild, Harris Semiconductor, Motorola, Intersil, Precision Monolithics, Raytheon and Texas Instruments, although National Semiconductor still offer the widest range of BiFET products, details of which can be found in their Linear Databook.

Very soon after National Semiconductor had announced the first BiFET products, RCA introduced their first BiMOS product, the economical CA3130 operational amplifier. This has some similarities to the BiFET amplifiers, but employs MOSFET transistors in the input stage rather than junction FET devices. RCA soon introduced further BiMOS devices, one of the best known type being the CA3140, which can be used as a pin-for-pin replacement for the 741 when a higher performance is required. More recently the CA080 series has been introduced as pin-for-pin replacements for the Texas Instruments series of TLO80 BiFET types.

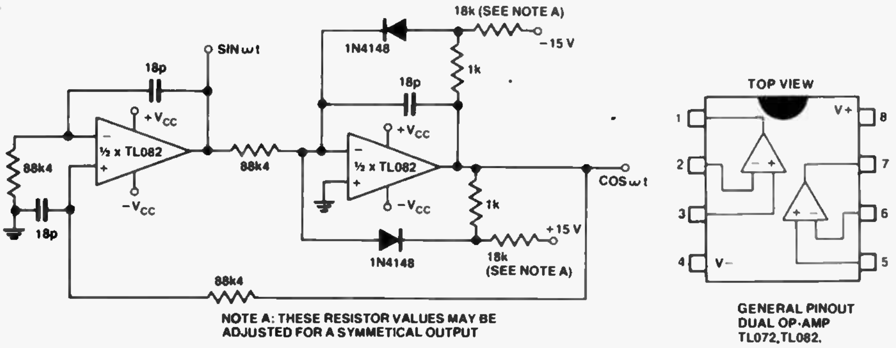


Figure 3. **Quadrature oscillator producing two outputs 90° out of phase, using a TL082 dual op-amp — pinout at right.**

professional engineer. Modern BiFET and BiMOS devices provide a very high input impedance with relatively good stability and temperature performance — although the input impedance of any of these devices at 25°C is much greater than over the full temperature range.

In general BiFET and BiMOS economical devices offer a comparable performance. If anything, BiMOS devices tend to offer the lower input bias currents and BiFET products the lower noise levels. However, premium devices of both types are available with performances far above the average for the type of device concerned.

Half-Hertz oscillator

Figure 2 shows the use of the economical TL081 device in a simple 0.5 Hz square wave oscillator. The TL081 is a single operational amplifier in a dual-in-line package with the connections shown in Figure 2; the pin connections are the same as those of the well-known 741 devices, internal frequency compensation being employed so that no external compensating capacitor is required. External offset adjustment can be made when required by means of an external variable resistor. The TL071 is a similar low-noise device with the same connections, and is quite suitable for use in this circuit, but its low-noise characteristics are not needed. The TL061 is a low-power device with the same connections.

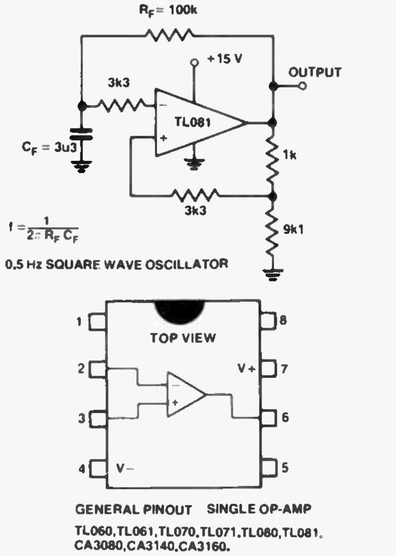
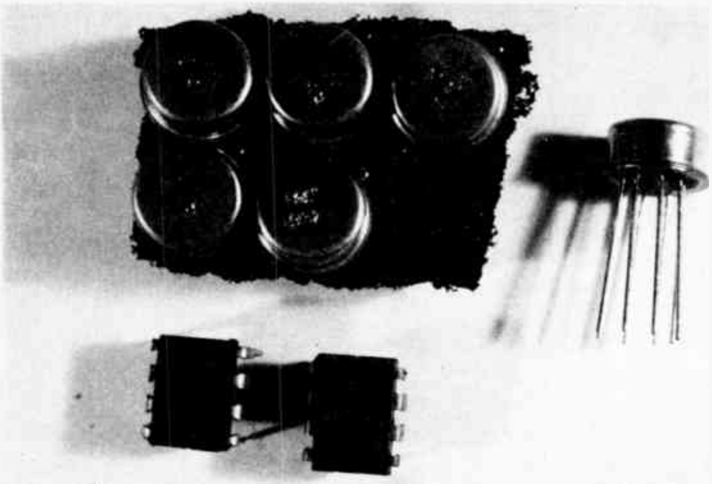


Figure 2. **Half-Hertz oscillator using a TL081 — pinout below.**



Modern BiMOS and BiFET op-amps come in both can and DIL packages.

The frequency of oscillation of the Figure 2 circuit is given by $f = 1/(2\pi R_F C_F)$, or about 0.5 Hz with the values shown. The high input impedance of the circuit enables a relatively high value of feedback resistor, R_F, to be employed, so the value of C_F can be reasonably small for a given frequency of operation. About nine-tenths of the output voltage is fed back to the non-inverting input to provide positive feedback to maintain oscillation. The capacitor C_F charges and discharges through R_F according to whether the state of the output voltage is 'high' or 'low' at the time concerned.

The circuit of Figure 2 generates square waves which are approximately symmetrical. However, if a circuit which generates waves with an unequal mark-to-space ratio is required, it is only necessary to connect a resistor of perhaps 10k to 50k in series with a diode across R_F. The direction in which the diode is connected determines whether the output spends the greater part of its time in the 'high' or in the 'low' state.

100 kHz oscillator

Figure 3 shows the circuit of a 100 kHz oscillator providing two outputs which are 90° out of phase with each other. Although the TL081 is perfectly satisfactory for use in this circuit, it is more convenient to use the dual TL082 device so that this one device is all that is needed. The connections of the 8-pin dual-in-line TL082 device are shown in Figure 3; it employs internal frequency compensation, but has no external offset adjustment facilities. ▶

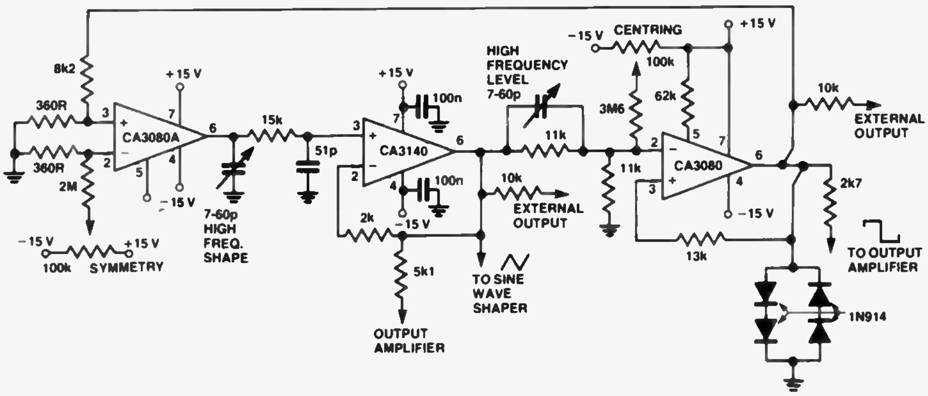


Figure 4. **Function generator circuit. Sourcing or sinking current from pin 5 of the left hand CA3080 will vary the frequency.**

BiMOS generator

A function generator which produces square and triangular waveforms is shown in Figure 4. It employs a CA3140 BiMOS device together with a CA3080A and CA3080. A particular feature of this circuit is that a frequency range of one million to one can be obtained by the use of a single variable resistor, or alternatively by the use of an auxiliary sweeping signal.

A CA3130 device may be employed instead of the CA3140 shown, but in this case a frequency compensating capacitor (about 56p) must be connected between pins 1 and 8, since the CA3130 is not internally compensated. The CA3160, which does not require any external frequency compensation, is also suitable for use in this circuit.

The high frequency linearity of the ramp is adjusted by the 7-60p variable capacitor connected between the output of the CA3140 and the output CA3080 device. The triangular wave output level is determined by the four 1N914 level-limiting diodes in the output circuit and the network connected to pin 2 of the CA3080.

It is important to minimise lead length and parasitic coupling capacitance in this circuit by careful layout.

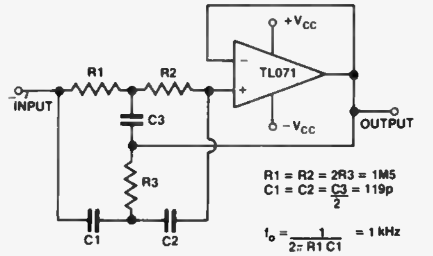


Figure 5. **Notch filter using a 'Twin-T' filter section on the input of a TL071 op-amp.**

Notch filter

The circuit of Figure 5 shows the use of a TL071 low-noise amplifier in a notch filter circuit. This is the normal 'twin-T' filter in the input circuit, in which one of the 'T' sections consists of R1, R2 and C3 and the other part of C1, C2 and R3. It is designed to reject signals of one particular frequency (the notch frequency), whilst passing signals of any other frequency virtually unattenuated.

For optimum performance, when a sharp notch in the frequency response is required, the components should have matched values (to within 1% or 2%). When the values shown are employed, the notch frequency occurs at approximately 1 kHz. An advantage of using a high input impedance device such as the TL071 is that relatively large values may be employed for R1, R2 and R3 and,

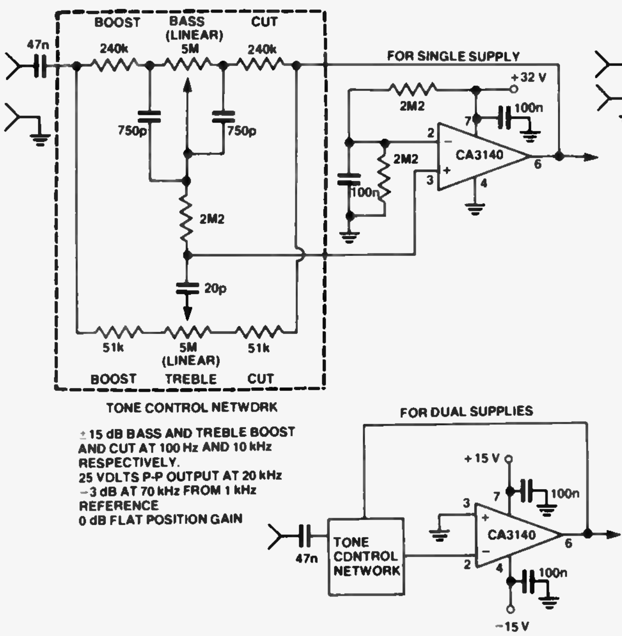


Figure 6. **Baxandall type tone control circuitry, with unity gain (flat position).**

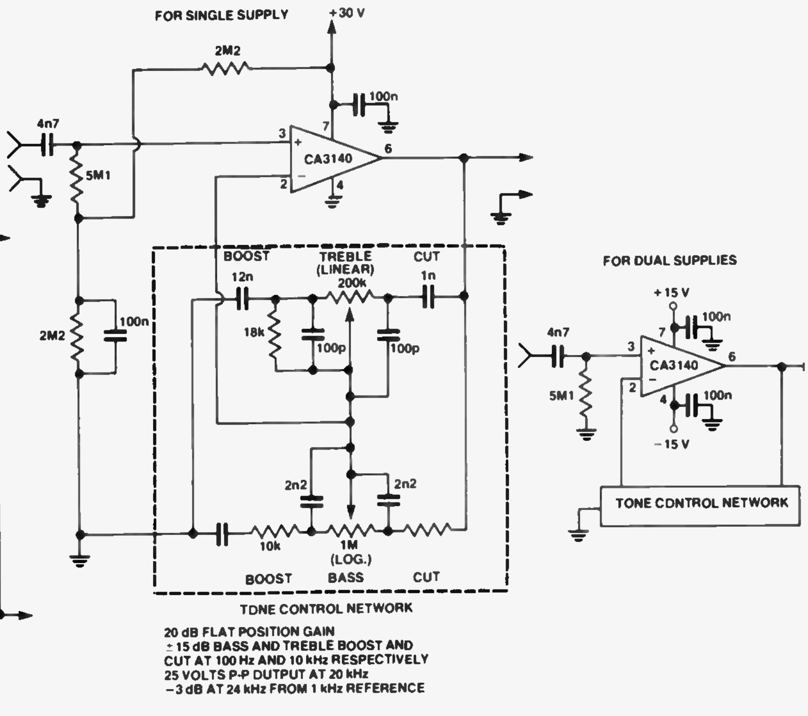


Figure 7. **Tone control circuit with 20 dB of gain, flat position.**

therefore, for any given frequency, C1, C2 and C3 can have a relatively low value. Large value, close tolerance capacitors are expensive, so the ability to employ devices of low value is important.

Tone controls

Two tone control circuits using the CA3140 are shown in Figures 6 and 7. Figure 6 is of the Baxandall type, which provides a gain of unity at the mid-frequencies and uses standard linear potentiometers. The high input impedance of the CA3140 enables low-value (and therefore cheap) capacitors to be employed in a circuit which has an impedance great enough to avoid excessive loading of the stage feeding this circuit.

Bass/treble boost or cut are about ±15 dB at 100 Hz and 10 kHz respectively. Full peak-to-peak audio output is available up to at least 20 kHz, since the CA3140 has a relatively high slew rate (about 7 V/us). The gain falls by about 3 dB at a frequency of around 70 kHz.

The circuit of Figure 7 provides similar boost and cut facilities, but the gain of this circuit is about eleven. The input impedance is basically equal to the resistor from pin 3 to ground.

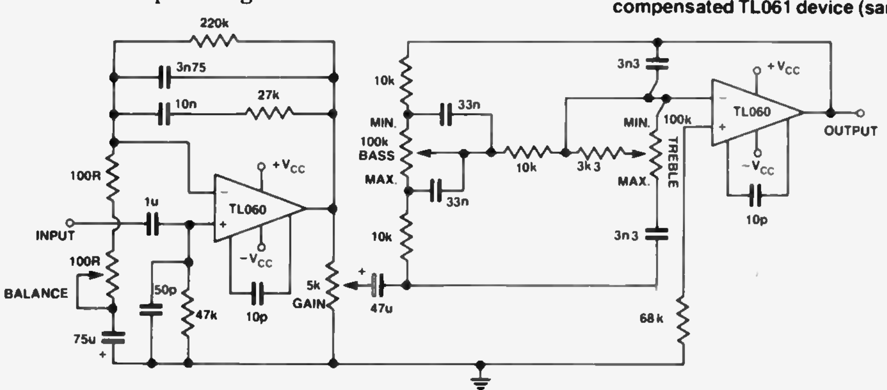


Figure 8. **A two-stage tone control circuit using TL060 devices. (Same pinout as TL070).**

A tone control circuit using the TL060 low-noise BiFET devices is shown in Figure 8. The TL060 is not internally compensated and therefore requires the 10p external frequency compensation capacitor shown connected in the circuit of each device. Similar circuits can, of course, be made using the TL080 devices at the expense of a higher power level. A further alternative is the use of TL066 programmable BiFET device without any compensating capacitors, but with a suitable value of the programming resistor between pin 8 and the negative line (about 1k, depending on the trade-off between bandwidth and power consumption which is required). Figure 9 shows the response of the Figure 8 circuit.

Mic preamp

A moving-coil microphone preamplifier with tone control is shown in Figure 10. A TL061 low-power device which is internally compensated is employed in this circuit.

Distribution amp

The Texas Instruments series of BiFET devices is also available with four separate amplifiers in a single 14-pin dual-in-line package. Figure 11 shows the connections of the TL064 low-power BiFET quad amplifier, together with a circuit for an audio distribution amplifier using one of these quad devices. The input stage acts as an input buffer and the other three stages act as output buffers, so that no signal from output A finds its way into any of the other outputs.

The TL084 and the low-noise TL074 have the same pin connections (which are the same as those of the LM324 type of device), whereas the TL085 and the low-noise TL075 devices are quad types with connections similar to the RC4136. There is no TL065 at present. ▶

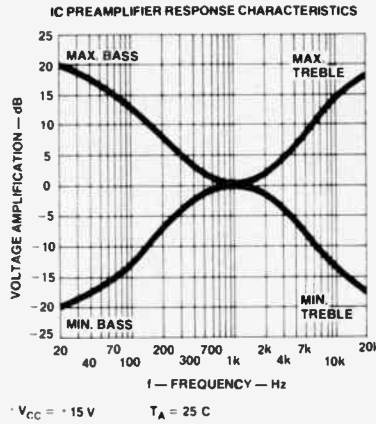


Figure 9. **Response characteristics of the Figure 8 circuit.**

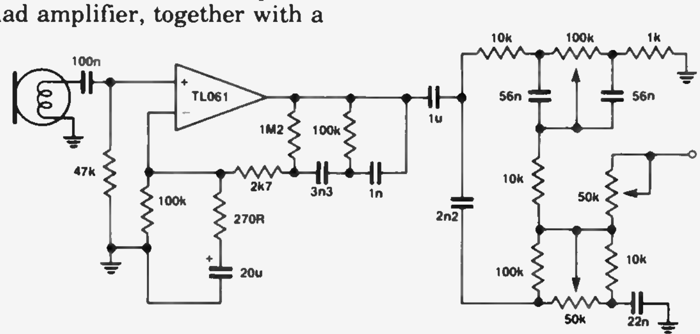


Figure 10. **Moving-coil mic preamp with tone controls, using an internally compensated TL061 device (same pinout as TL071).**

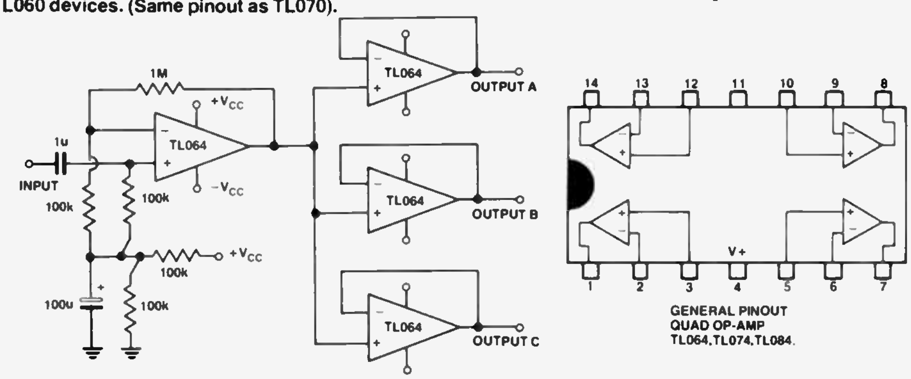


Figure 11. **An audio 'distribution' amplifier for 'slaving' several pieces of equipment from a single source. Pinout for the quad op-amp is shown at right.**

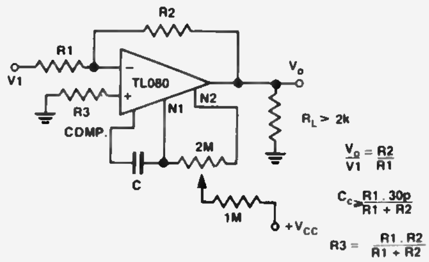


Figure 12. **Simple voltage-variable gain amp using the TL080.**

Variable gain

The simple circuit of Figure 12 is an amplifier which provides a variable gain set by the potentiometer. A TL080 device is employed, so the compensating capacitor C_c is required, since this device is not internally compensated.

Ice warning

The circuit of Figure 13 employs three of the four amplifiers of the TL084 device in an ice warning detector. It is especially suitable for use in vehicles to warn the driver when the temperature of the thermistor (placed outside the vehicle) falls below 0°C.

When the temperature of the thermistor falls, its resistance rises and the current flowing through the thermistor decreases. Thus the inverting input of the TL084 connected to this thermistor receives less current from the positive supply line and its output voltage tends to rise. This output voltage is fed to the TL084 output amplifier and produces a voltage across the LED, which lights, providing the required warning.

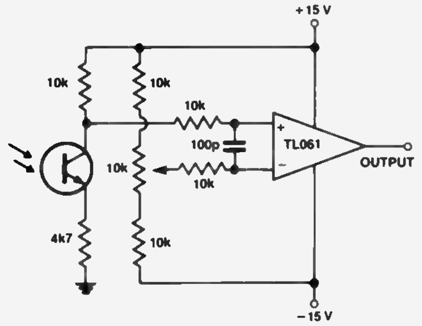


Figure 14. **Low-level light detector using FPT100 or similar phototransistor.**

Light detector

The circuit of Figure 14 is a low-level light detector preamplifier using the low-power TL061 device with a TIL601 or similar phototransistor. The variable resistor can be used to balance the output at any particular value of light level.

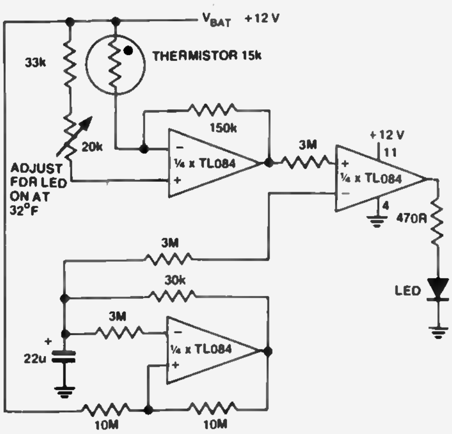


Figure 13. **An ice warning indicator.**

Sine shaper

The circuit shown in Figure 15 uses a CA3140 as a voltage follower device in combination with diodes from the CA3019 array to convert the triangular signal from a function generator into a sinewave output, which has typically less than 2% harmonic distortion.

The circuit is best adjusted using a distortion analyser, but a fairly good adjustment can be made by comparing its output signal on an oscilloscope with that from a good sinewave signal generator. The initial slope is adjustd by R1, followed by an adjustment of R2. The final slope is established by adjusting R3, thereby adding additional segments that are contributed by these diodes. Repetition of the adjustments may be necessary, since there is some interaction between the adjusting potentiometers.

Wien bridge

A CA3140 BiMOS amplifier is used in the circuit of Figure 16, together with a CA3019 diode array, to form a Wien bridge oscillator. The zener diode shunts the 75k feedback resistor and, as the output signal amplitude increases, the zener diode impedance rapidly decreases so as to produce more feedback, with a consequent reduction in gain. This action stabilises the output signal amplitude. This combination of a monolithic zener diode and the bridge rectifier tends to provide a zero temperature coefficient for this regulating system.

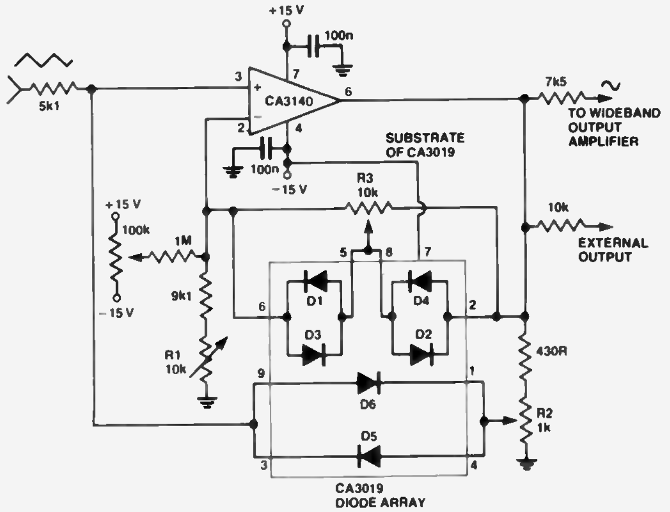


Figure 15. **A triangle-to-sine waveshaping circuit employing a CA3140 op-amp and a CA3019 diode array.**

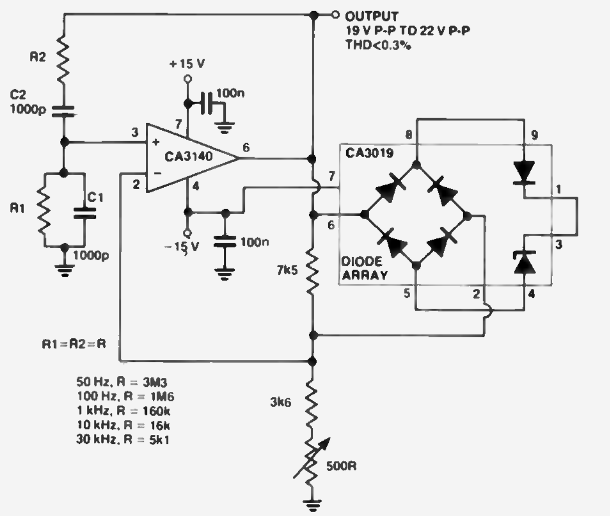


Figure 16. **A Wien bridge oscillator featuring amplitude stabilisation via the zener action from the CA3019 diode array.**

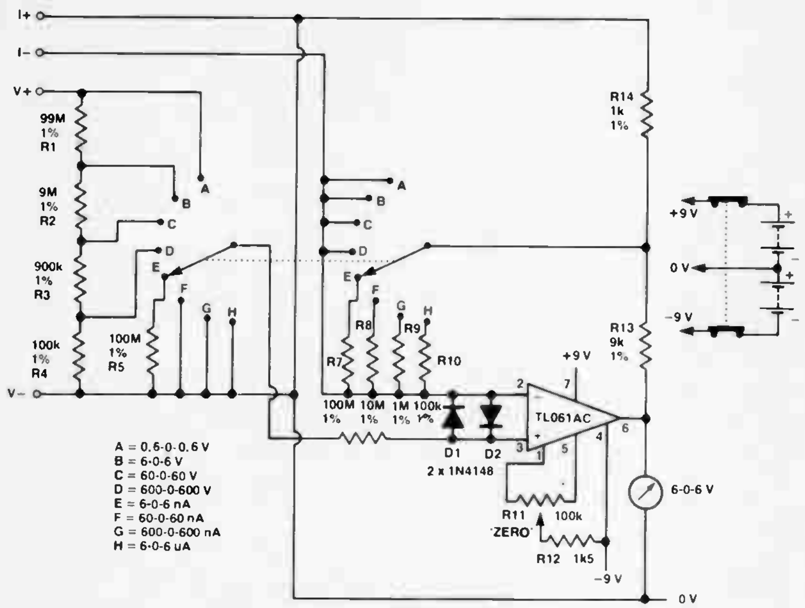


Figure 17. **A multi-range voltmeter with high impedance input plus multi-range low-current meter.**

As the output circuit contains no RC time constant, there is no lower frequency limit for operation. If C1 = C2 = 1u (polycarbonate) and R1 = R2 = 22M, the operating frequency can be about 0.007 Hz. At high frequencies, as the frequency is increased the amplitude of the signal must be reduced to prevent slew rate limiting from taking place. An output frequency of about 180 kHz will reach a slew rate of about 9 V/us when the output voltage amplitude is about 16 V peak-to-peak.

Meter

The high input impedance of BiFET and BiMOS devices has led to their use in many voltmeters of high input resistance and also in meters to measure very small currents.

The circuit of Figure 17 was designed by Texas Instruments for the measurement of voltages in the range ±0.6 V to ±600 V, where the source resistance may be quite high, and to measure currents from 6 nA to 6 uA. The instrument was required to accept inputs of either polarity and be inexpensive, robust and reliable. It also had to have a long battery life, so a TL061 low-power operational amplifier device was selected. An inexpensive centre zero meter is considerably cheaper than a liquid crystal display and would provide adequate accuracy for the purpose.

When the switch is in one of the positions A to D inclusive, the instrument is set for the measurement of voltages. The amplifier has a non-inverting gain of 10 and range selection is achieved by a simple potential divider network with a fixed input impedance of 1000 megohm. A panel-mounted 'centre zero' control is included in the circuit to facilitate corrections for the mechanical movement of the meter zero and for the change in the operational amplifier input voltage offset (for example, with temperature).

In the current measuring mode of switch positions E to H inclusive, the amplifier operates as a current-to-voltage converter. For the most sensitive range of 6 nA, a transimpedance of 1 Gigaohm is required to produce a full-scale deflection of the meter. Rather than use a resistor of such a high value, a resistance multiplier arrangement was devised with a 100M feedback resistor for the most sensitive range.

The two diodes across the input of the operational amplifier in conjunction with R6 provide protection against any gross overloading of the instrument. A suitable arrangement incorporating a fullwave rectifier into this circuit would allow alternating input signals to be measured, but arrangements would have to be made to allow for frequency roll-off of the response at high frequencies.

3 pA meter

A CA3160 and a CA3140 are used in the circuit of Figure 18 to construct a picoammeter with ±3 pA full scale deflection (one picoamp = 10^{-12} amps). Pins 2 and 4 of the CA3160 are connected to ground, so the input pin 3 between them is effectively 'guarded'. If slight leakage resistance is present between terminals 3 and 2 or 3 and 4, there would be zero voltage across this leakage resistance and this would reduce the leakage current by a large factor.

It is preferable to operate the CA3160 with its output pin 6 near the ground potential, so as to reduce the dissipation by reducing the device supply current. The CA3140 serves as a x100 gain stage to provide the required plus and minus output voltage swing for the meter and feedback network. A 100:1 voltage divider network consisting of a 9k9 resistor in series with a 100 ohm resistor sets the voltage at the 10 kMohm resistor to ±30 mV full-scale deflection. This 30 mV signal results from ±3 V appearing at the top of the voltage divider network, which also drives the meter circuitry.

It is possible to switch the 9k9 and 100 ohm network in the output circuit so that current ranges from 3 pA to 1 nA can be handled using the single 10kM resistor.

The writer has seen circuits using BiMOS devices published for use in measuring currents down to 100 femtoamps (0.1 pA), but obviously extreme care is required to ensure the insulation is adequate when such small currents are being measured. ▶

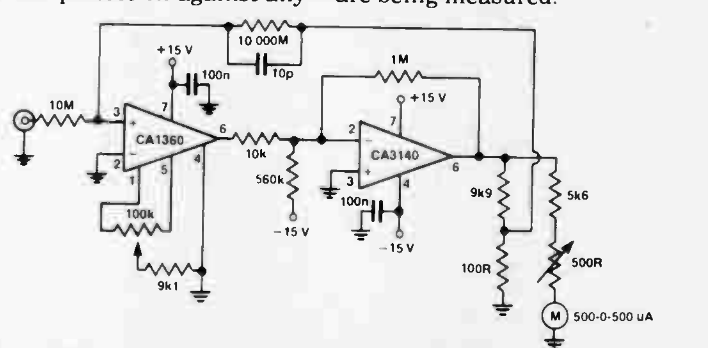


Figure 18. **This circuit will measure *very* low currents — full-scale deflection is ± three picoamps!**

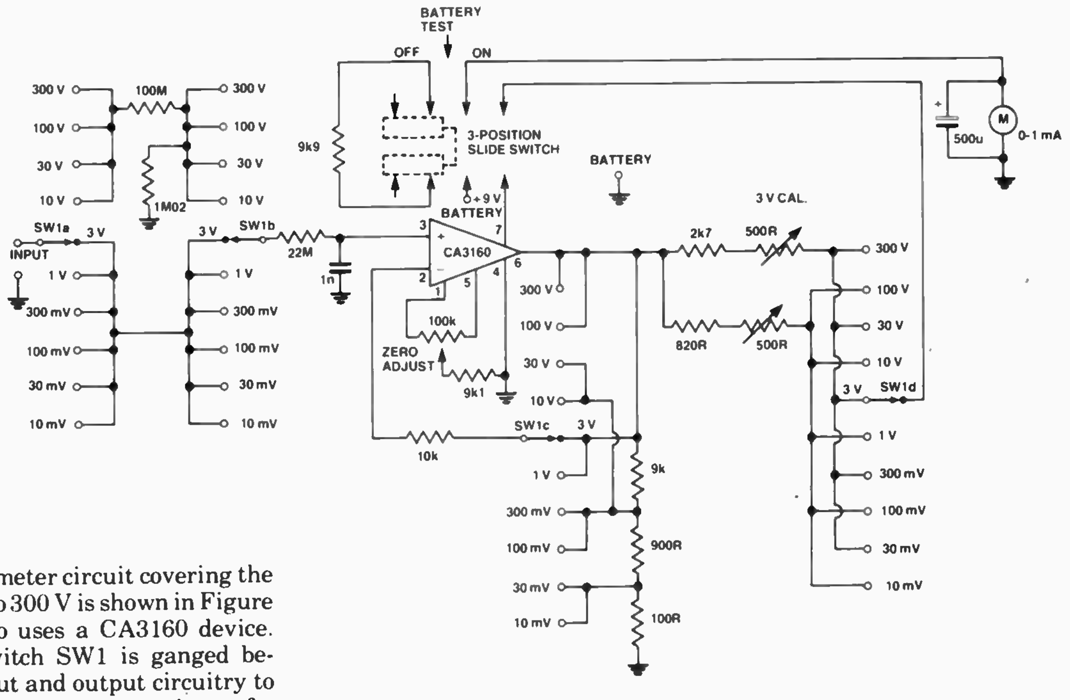


Figure 19. **Example of a multi-range voltmeter measuring from 10 mV to 300 V.**

Voltmeter

A further voltmeter circuit covering the range 10 mV to 300 V is shown in Figure 19, which also uses a CA3160 device. The range switch SW1 is ganged between the input and output circuitry to enable the proper output voltage for feedback to terminal 2 through the 10k resistor to be selected.

This circuit is powered by a single 8.4 V mercury battery, the power supply current being somewhat less than 500 uA plus the meter current required to indicate a given voltage. Thus the supply current rises to about 1.5 mA at full-scale deflection.

Any readers who experience problems in obtaining a CA3160 may use a CA3130 with a frequency compensation capacitor of about 56p between pins 1 and 8.

The aim of this article has not been to introduce readers to all the latest BiFET and BiMOS devices (of which there are large numbers), but rather to give an indication of the wide selection of circuits that can be made with just a few of the standard types of device which are readily available. ●

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A number of useful designs include a speech processor, interference filters and a simple CB radio receiver. Stripboard layouts, wiring diagrams and notes on construction are provided.

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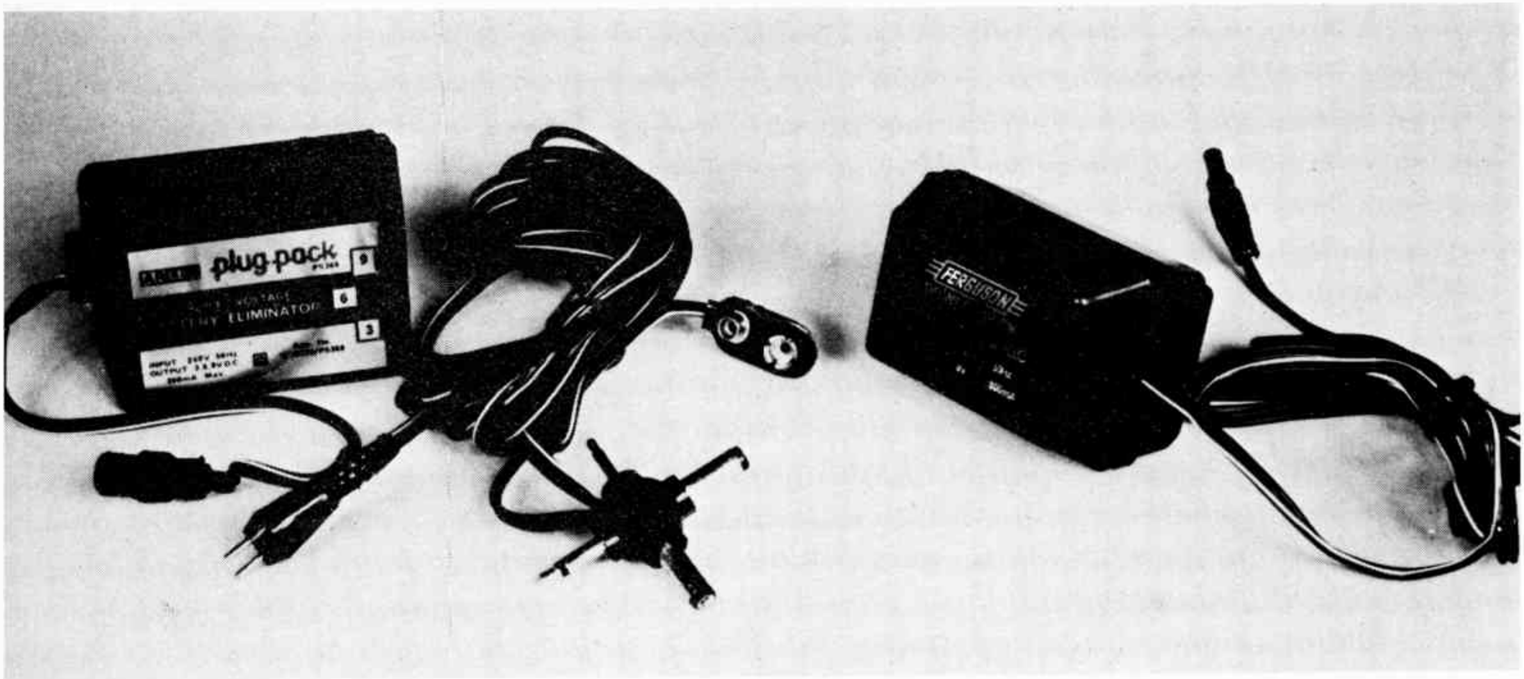
"Plug Pack" battery eliminators

Commercially available battery eliminators can often be used in place of specially built and costly power supplies.



About the most convenient and least expensive mains power supplies suitable for powering our simple battery-operated projects are these "Plug Pack" battery eliminators. At left is the PS369 from A&R Soanar – this unit gives three, selectable, voltages and comes with a range of output connectors, a very versatile unit. At right is the PPA9-DC from Ferguson Transformers Pty Ltd, which delivers 9 Vdc at up to 3 VA.

LOOKING THROUGH our pages each month, as well as the 'Project Electronics' and 'Simple Projects' series, reveals a myriad of simple projects, most of which use a single nine volt power supply. In fact we try to design all our simple projects to use the common No. 216 transistor radio battery.

After building a few of these projects you may have thought a simple mains power supply would be a good idea to reduce the turnover of batteries, which in some cases are the most expensive item in the circuit! The power supply would not have to be complicated, good regulation being unnecessary for most of these projects, but by the time you bought a transformer, diode bridge, filter capacitor, box and leads you would have spent the best part of $15. That's about 20 batteries worth at today's prices – and chances are if you're a newcomer to electronics, building our simpler projects, you shouldn't be thinking of building mains powered equipment for a while. We'd like to keep our enthusiastic readers – not kill them off !

The two largest Australian transformer manufacturers, Ferguson Transformers Pty Ltd and A & R Soanar Pty Ltd, have come to the rescue with a range of "Plug Packs". These consist of a small plastic case containing a small transformer, diode rectifier and filter capacitor. The case has a moulded-in mains plug (generally two pins) allowing the whole unit to be plugged into a standard 240 Vac outlet. A twin-lead and low voltage connector provides output connection.

Originally designed to power portable radios, calculators and cassette recorders, they are available in a range of voltages and power ratings, and are ideally suited to powering our simple projects.

Ferguson plug packs are locally made and available in 3, 4½, 6, 7½, and 9 volts at a rated output of 3 VA (watts) or 500 mA, whichever is the greater. They are terminated in a standard connector found on most portable equipment. The nine volt version (PPA 9-DC) is probably the most useful, to project constructors, of the Ferguson range.

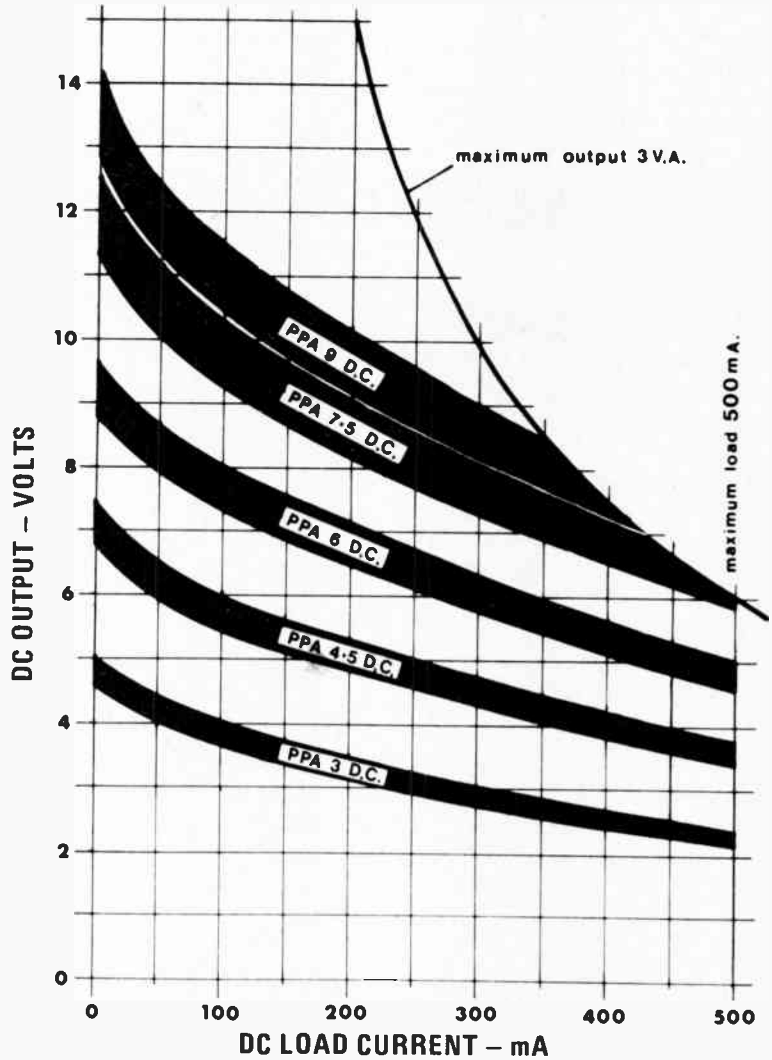
A&R Soanar have come up with a novel idea in one of their Plug Packs. Rather than have a range of different voltage packs they use a single pack with a tapped transformer to give three, six or nine volts at an output of 300 mA. These have a multi connector with two sizes of jack plugs, two sizes of low voltage coaxial type connector and a battery clip for equipment without a remote power socket. Priced at around $10 it is more expensive than the Ferguson range and has a lower output current but offers greater flexibility. You need to be careful though when using them not to short out the unused connectors.

Output voltages are rated at a specific current. As the supplies are unregulated the output voltage will be higher at low currents and fall below the rated output at high currents. This poor regulation will not effect most battery equipment or the operation of our projects, in general.

The graph here gives the output voltage for specific currents for each of the Ferguson range. The maximum current for the lower voltages is limited at 500 ma and at 3 VA for the higher voltage packs. The 3 VA curve can be seen at the top right of the graph. If an output of say 12 V dc at 40 mA is required it can be seen that a 9V Plug Pack will do nicely, however this drops to about 8 V at full load.

Plug Packs can also be used for charging Nickel-Cadmium (NiCad) batteries if the right pack is selected for the capacity and voltage of the batteries to be charged. Nicads must not be charged with a current greater than one tenth of their amp-hour capacity. The PPA9-DC will supply 400 mA maximum current and can be used to charge a 12 V 4 AH Nicad pack. Smaller NiCad packs will require a series resistor to limit the maximum charging current. Penlite size cells requiring only 40 mA charging current will require about a 330 ohm series resistor.

Plug Packs can be obtained from many of the suppliers who advertise in the magazine, and you should have little difficulty obtaining what you require ●



Curves showing the output voltage ranges and regulation for the various Ferguson model "Power Pack" battery eliminators. All deliver their rated output voltage at a current of 300 mA. They will give a higher voltage at low load currents and lower output voltage at high load currents, but note that the 3 VA (that's 3 watts generally in the applications we suggest) must not be exceeded for the PPA9-DC or PPA7.5-DC units. The width of each curve indicates the maximum and minimum output voltages for each unit.

Light and power from dc supplies

Generating mains-independent light and power from batteries is fraught with many unrealised difficulties. Whether you want dc back-up to operate equipment when the mains goes 'off-the-air' or a wholly independent 240 Vac supply, you should know the problems up front.

Roger Harrison

THAT'S THE TROUBLE with Electricity Commissions — they've insidiously crept into our lives and made us quite dependent on them. For those occasions when we cannot avail ourselves of their 'services', we have to rely on other sources to provide light and power. The old kerosene pressure lamp has its advantages — and disadvantages — but how on earth do you keep a disk drive running when the ac mains 'browns out'? As storage batteries are ubiquitous, the 12 V car battery in particular, it's natural that we turn to them to provide back-up and mains-independent supplies.

Back-up supplies

For equipment designed to be powered directly from a nominal 12 Vdc source or from either 12 Vdc or 240 Vac, back-up supplies are employed to maintain continuity of supply, the battery being kept charged from the mains, but the battery acts to maintain power supply to the equipment in the event of mains failure. This sort of system is commonly installed with burglar alarms, amateur radio repeaters and geophysical monitoring equipment, for example.

The 'power budget' of such systems is carefully considered to provide maximum service period from the battery supply when mains is unavailable. Hence a single 12 V storage battery — generally a low maintenance type — is employed. Let's learn a bit about lead-acid batteries first.

The fully-charged, no-load terminal voltage of a lead-acid cell is between 2.3-2.4 volts. This drops under load to about 2.0-2.2 volts. When discharged, the cell voltage is typically 1.85 volts. The amp-hour capacity is determined from a 10-hour discharge rate. The current required to discharge the battery to its end-point voltage of 1.85 V/cell is multiplied by this time; e.g: a 40 Ah battery will provide four amps for 10 hours before requiring recharge. Note however that the amp-hour capacity varies with the discharge current. The same battery discharged at a rate of 10 amps will not last four hours; on the other hand if it is discharged at 1 amp it will last somewhat longer than 40 hours. The typical discharge characteristics of a (nominal) 12 V battery are shown in Figure 1.

The ideal initial charging current for the fully discharged battery (cell voltage under 2.0 V) should be about 20 amps per 100 amp-hours of capacity (i.e: 8 amps for a 40 Ah battery). Once the electrolyte begins to gas rapidly, the terminal voltage will be around 13.8 volts and rising rapidly. At this point, the charging current should be reduced to somewhere between 4-8 amps per 100 Ah until charging is complete.

At the end of charging, terminal voltage may rise to about 15.6 volts or more, but this decreases slowly after the charger is removed, the terminal voltage then usually reading around 14.0 to 14.4 volts (see Figure 2).

Back-up supplies are generally of the 'trickle-charge' type or the 'battery condition' sensing type. Two good examples are ETI projects 597 — Emergency Lighting Unit (December 1980) — and 1503 — Intelligent Battery Charger (August 1981). The ETI-597 trickle charges a 12 V battery when the mains is on and provides automatic switchover when the power drops out. It's cheap and simple, but needs to be *used* for the batteries to stay in condition so that they deliver their rated capacity when

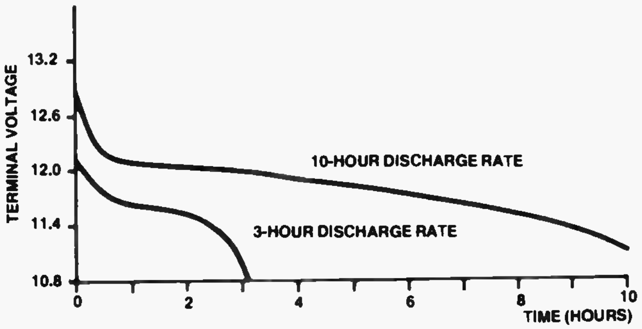


Figure 1. **Typical discharge characteristics of a 12 V (nominal) lead-acid battery.**

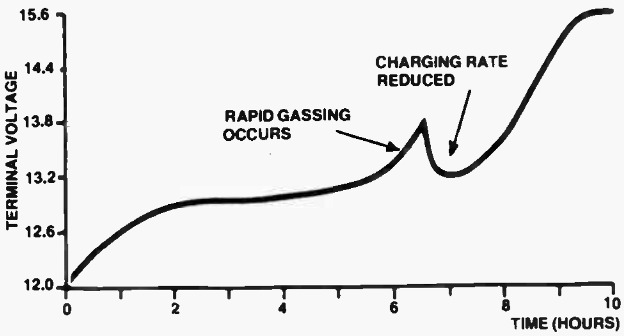


Figure 2. **Charging characteristics of a 12 V (nominal) lead-acid battery. The 'kink' in the curve near six hours is explained in the text.**

needed. Back-up supplies of this sort are only practical where the load on the supply is not too heavy — generally 20 W or so.

To drive a heavier load, upwards of 50 W for example, it's best to power the equipment from the battery *all the time* and have a charger which senses the battery terminal voltage, charging the battery when the terminal voltage falls to a preset level and turning off when the terminal voltage rises to the desired operating level again. This is what the ETI-1503 does. There is a slight element of luck involved as to how charged the battery will be at any one time, but the lower limit is usually set so that the equipment will operate for a specified period. The ETI-1503 can be used with batteries with a capacity up to 100 Ah. Such a battery can drive a 10 A load at the 10-hour discharge rate — which effectively means it's a good back-up supply for equipment with a power budget of up to 120 W mean consumption. This means that actual consumption can be greater than that from time to time provided that consumption falls below the mean level for an equivalent period. An amateur VHF or UHF repeater is a good example. Whilst 'listening' only — no stations active on the input channel — consumption is quite low. When 'activated' by a station or stations, the repeater spends most of its time transmitting, and consumption can be four to ten times that during inactive periods, depending on the power output of the transmitter employed in the repeater.

As stated earlier, the major consideration with back-up supplies is the power budget of the equipment being supplied. If you anticipate the necessity of operating the equipment for periods exceeding, say, eight hours, then a battery of adequate ampere-hour capacity needs to be used. It is always prudent to choose a battery with 20-50% more capacity than strictly necessary.

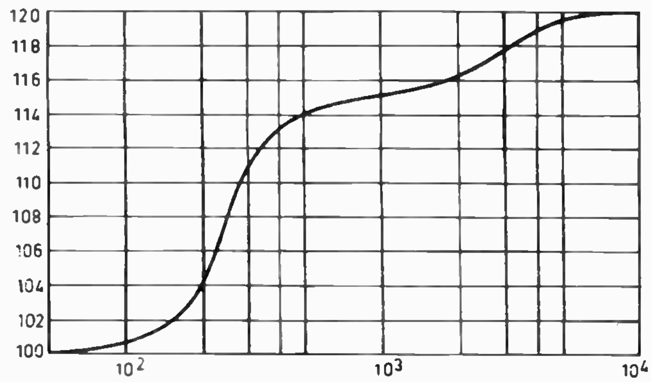


Figure 3. **The light output of a fluorescent tube increases with increasing supply frequency in the manner shown in this graph. The property is exploited in dc-ac square wave inverters for lighting.**

dc-ac inverters

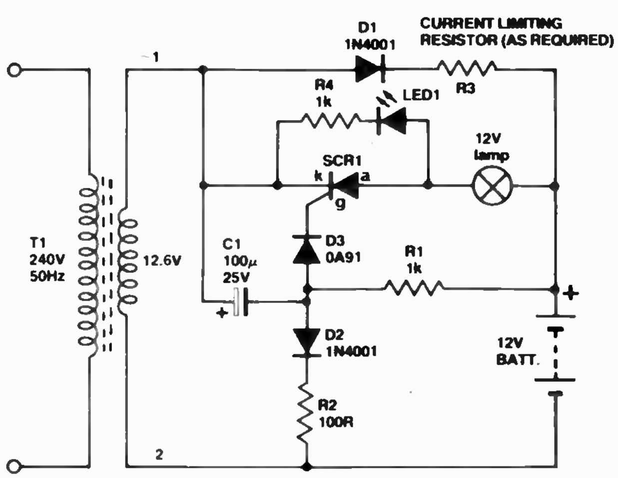
Like storage batteries, 240 Vac mains-operated equipment is ubiquitous! The huge variety of products have been designed to be *convenient*, thus making themselves *necessary*. Or so it seems. Why on earth anyone would want to take an electric razor on a camping expedition and expect to power it from an ersatz 240 Vac supply is beyond this writer — but then I haven't had a shave in more than 15 years except when my appendix was removed and then they didn't shave my face!

There are two common approaches to providing 50 Hz ac power for mains operated appliances: provide square wave drive of the appropriate amplitude, or derive a sinewave (or pseudo sinewave) supply of appropriate amplitude. Both are fraught with hidden difficulties. If you want any substantial amount of power output — like 200 W — you're in hot water — and probably unable to boil a billy, to boot!

A square wave dc-ac inverter has the advantage of simplicity and efficiency — depending somewhat on the design. Inverters generally take two forms: *self-excited,* usually employing a feedback winding on the transformer, and *driven,* where an oscillator drives a switching circuit, generally with transformer output. Where the precise frequency of the ac output is unimportant, self-excited inverters are employed. Where a stable 50 Hz output is required, a driven inverter is necessary.

Lighting is one area where self-excited dc-ac inverters find application. The common tungsten filament incandescent light globe is a poor choice for lighting where a dc supply is employed. They have an efficiency of less than a fifth of that of a fluorescent light of the same power rating — viz: around 12 lumens/watt for the tungsten filament lamp versus better than 60 lumens/watt for a fluorescent tube. A 20 W fluorescent tube would provide as much light output as a 100 W incandescent globe! Those figures are based on 50 Hz ac supply. Fluorescent tubes *actually improve in efficiency* when driven from a higher frequency supply. Figure 3 shows how the light output of a fluorescent tube increases with increasing supply frequency. Driving the tube from a supply frequency of 10 kHz or more will result in a 20% increase in light output.

The circuit of a self-excited inverter driving a fluorescent tube is shown in Figure 4. This is taken from Project ETI-516 of November 1972. It ran at around 2 kHz and employed a ferrite-cored transformer. Consumption was 2.5 amps. An incandescent globe to pro- ▶



Circuit of the ETI-597 'Emergency Lighting Unit', a simple back-up supply that can be used for other than lighting applications.

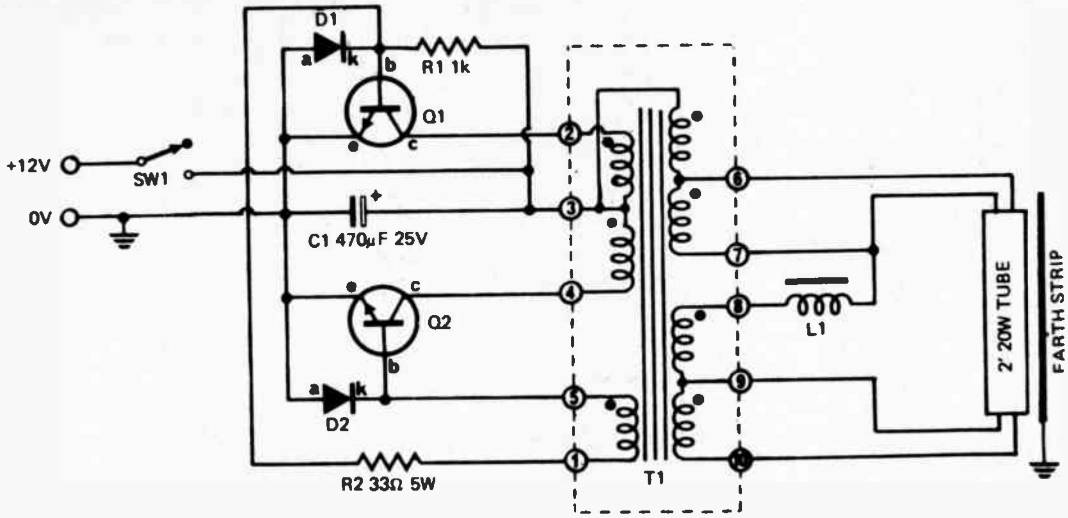


Figure 4. **ETI project 516 (Nov. '72!) employed a self-excited dc-ac square wave inverter operating at 2 kHz to drive a 20 W fluorescent tube — an efficient solution to providing light from a dc supply.**

vide a similar light output would draw around 10 amps! Such inverters have one drawback — the transformer core 'sings' owing to the magnetostrictive forces on the core pieces (which generally come in two pieces). That can be solved in two ways — put the inverter in a 'soundproof' box or operate the inverter at a frequency above audibility. The first solution was employed with the ETI-516 inverter, but is inevitably only partially successful (though often acceptable).

When it comes to powering 240 Vac-operated equipment or appliances a number of considerations have to be looked at. First, will the equipment operate from a square wave supply? Many appliances employing an ac or ac/dc motor will operate quite happily from a square wave supply. One of ETI's correspondents employed battery back-up for his computer's disk drives, supplying these with 240 V, 50 Hz square wave ac from a driven inverter. The general arrangement is shown in Figure 5. A 100 Hz oscillator drives a flip-flop, which drives a pair of HEXFETs connected in push-pull across the secondary of a toroidal transformer. Battery supply was 24 V. The transformer is operated 'back-to-front' here, where input is applied to the secondary and the load connected across the primary. Toroidal transformers perform much better in this application than conventional types as core losses are lower and primary-to-secondary coupling is generally better. Some losses are involved, the saturation voltage of the HEXFETs generally being the greatest source. Hence the use of a 20-0-20 V winding and not a 24-0-24 V winding.

The saturation voltage loss in switching devices driving a transformer is an important consideration. One or two volts lost from a 24 V supply represents only about 4% to 8% loss, but at 12 V it's twice that! Any further losses only magnify the problem.

A square wave ac supply is inherently rich in harmonics. These can play havoc with audio and digital equipment and it's often difficult to suppress interference generated by the supply. Then again, some equipment — particularly anything containing a transformer and rectifier, will produce entirely different performance from when it's operated from a sine wave supply. The problem arises because the peak and RMS values of a square wave are the same, whereas the peak/RMS ratio for a sinewave is 1.414. To deliver the same *work value* as a sine wave supply, the peak output voltage of a square wave dc-ac inverter is generally set at 240 V. When driving a motor or resistive load, the square wave supply will deliver the same amount of power as a sine wave supply; i.e: the same amount of *work* will be done (all else being equal). But, where the load or equipment expects a peak voltage of 340 V (as we have with the ordinary mains), then a square wave supply of a nominal 240 V output will not 'deliver the goods' as its peak voltage is only 240 V.

So much for that; let's look at sine-wave dc-ac inverters. At this stage, I recommend you read the letter from reader Barry Brown in the accompanying panel.

Requests of a similar nature arrive quite commonly, though Mr Brown's is a little unusual compared to many we receive. Where Mr Brown suggests a dc-ac inverter to operate from a 24 V or 32 V supply, many readers ask for a

Dear Sir,

Despite the financial burden of component costs for an ever-expanding range of new projects, please find enclosed my cheque for subscription renewal. Would you consider a three-to-six-month moratorium on new projects to enable those of us who are more enthusiastic about starting the new than finishing the old to clear some of the backlog?

No?

Well, to an area with possibly more appeal — that of small domestic power supplies. With most homesteads using 32 Vdc power for lighting and almost all travellers using 24 or 12 Vdc power, a stable inverter producing 240 Vac at 50 Hz in the 500 W-1 kW range could be of enormous benefit to many people. With an increasing range of domestic electronic equipment becoming available, the only way for many people to enjoy these products is to crank over the, if available, 240 V diesel generator. This is great stuff during the day, especially with auto-start, but can be a bit distressing when you discover that your evening music selection is really a duo for harpsichord and Lister Diesel.

More ambitious, and probably a lot heavier: a larger inverter capable of starting a fridge/freezer would win many friends. Although these devices may have considerable losses in conversion, there are advantages in using converters for many applications. Not the least of these would be the lifetime supply of European carp likely to be donated by the Darling River United Naturalists Kangaroo Appreciation, Research and Development Society (DRUNKARDS). Such a society, although unformed, could be initiated at the drop of a cold tinny after publication of a suitable circuit.

Barry Brown,
Young NSW.

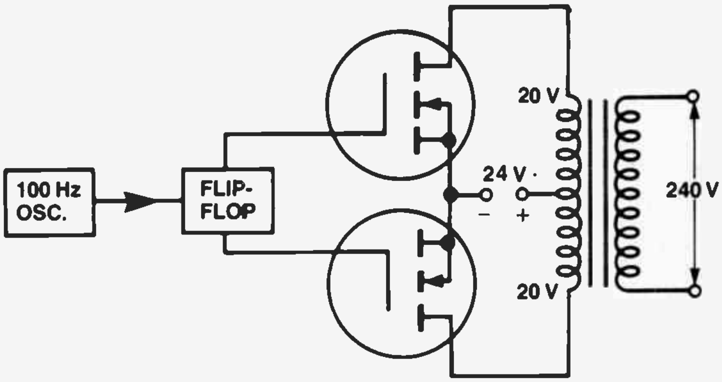


Figure 5. **Example of a 'driven' dc-ac square wave inverter with nominal 240 Vac output. This technique has been employed by one of our correspondents as a computer back-up supply.**

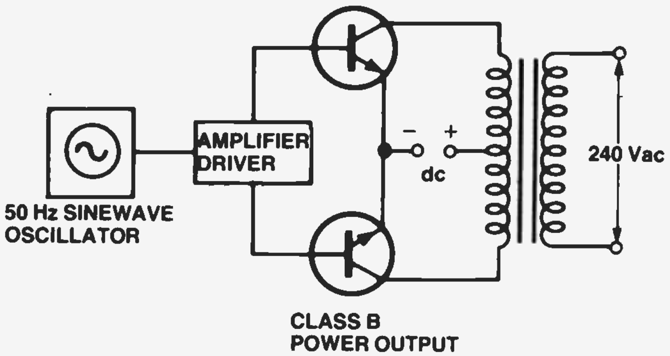


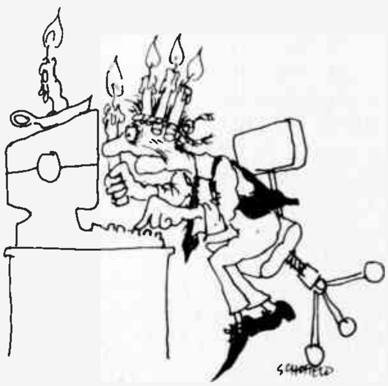
Figure 6. **Class B driven sinewave inverter technique for providing 240 Vac from a dc supply.**

1 kW or similarly rated inverter to run from a 12 V battery. The latter is impractical, for the following reasons.

Consider this: a sinewave dc-ac inverter needs to be of the driven type. Hence it generally consists of an oscillator driving a class B power amplifier — usually a push-pull type. The theoretical maximum efficiency obtainable with a class B power amplifier is 78%. With losses and power consumption of drive circuitry taken into account the dc power input to ac power output efficiency of an inverter of this type is generally around 65-70%. Thus a 1 kW dc-ac inverter to run from a 12 V battery *would draw in excess of 120 amps* at full load! Few batteries available would supply that sort of current for long. With currents of that magnitude, special arrangements have to be made for primary circuit conductors. A resistance of 5 milliohms (0.005 ohms) will result in a power loss of more than 70 watts. Then again, special consideration has to be given to heat dissipation in the power output stage. The devices used would dissipate something over 400 W at peak load. No load dissipation would probably be in the vicinity of 40-50 W, which is no mean amount to get rid of.

Apart from the weight of a heatsink, consider the weight of a 1 kVA (or 1000 W) transformer (assuming a single transformer is used). We'll leave the expense to your imagination.

The problems are reduced somewhat when a much higher dc supply voltage is available. However, in the latter case other techniques of dc to ac conversion present themselves — but that should be the subject of another article as it's a whole new ballgame.



Where a 12 V battery supply only is available, there is a practical limit to the maximum power of a dc-ac inverter, and that's probably around 300 W output. At typical efficiencies, the dc input power is around 450 W, or close to 35-40 amps current from the battery.

As you would already appreciate, this brings its own special problems. A battery to supply that sort of power for any appreciable or worthwhile period would need to have a considerable ampere-hour capacity. Your typical 40-60 Ah car battery would barely deliver an hour's worth of power. If the inverter is installed within the vehicle, or close by, and you are willing to keep the engine running during operation, then the battery will deliver the goods for quite a period, provided you can 'set' the throttle to suit so that battery charge is maintained. At this stage, I might point out that an alternator coupled to the motor would provide a more efficient energy conversion.

To gain, say, four to six hours of operation for a 300 W inverter, you would need a battery system of more than 200 Ah capacity.

A more practicable power level for a sinewave dc-ac inverter would be around 120 W. Such an inverter would pull 12 to 15 amps from the battery, a much more manageable figure.

Having seen the primary side of the problem, let's consider the secondary side — the load. How many appliances do you have rated at less than 300 watts? Very few. The humble electric kettle is rated from 1 kW to 2.4 kW. Monochrome TV sets, particularly portables, may only consume 100 W, but a colour TV may draw three times that or more. A 'low power' (say, 30 W/ch.) domestic hi-fi will draw around 100 W, depending on how much equipment is in use and how loud you like it. Anything more ambitious has a proportionately larger consumption. A 300 W dc-ac inverter is best considered where the full output is only required intermittently.

Conclusion

As can be seen, many factors have to be taken into account when considering obtaining light and power from a battery supply — whether it be in a back-up application, for lighting or 240 Vac substitution. The ubiquitous 12 V battery is not up to the job in some instances — in which case higher voltage dc systems are better considered.●

Integrated switching regulator

Switching power supplies are replacing linear supplies in many applications because of their higher efficiency, reduced size and generally lower cost. National Semiconductor has introduced a switching regulator building block, the LH1605.

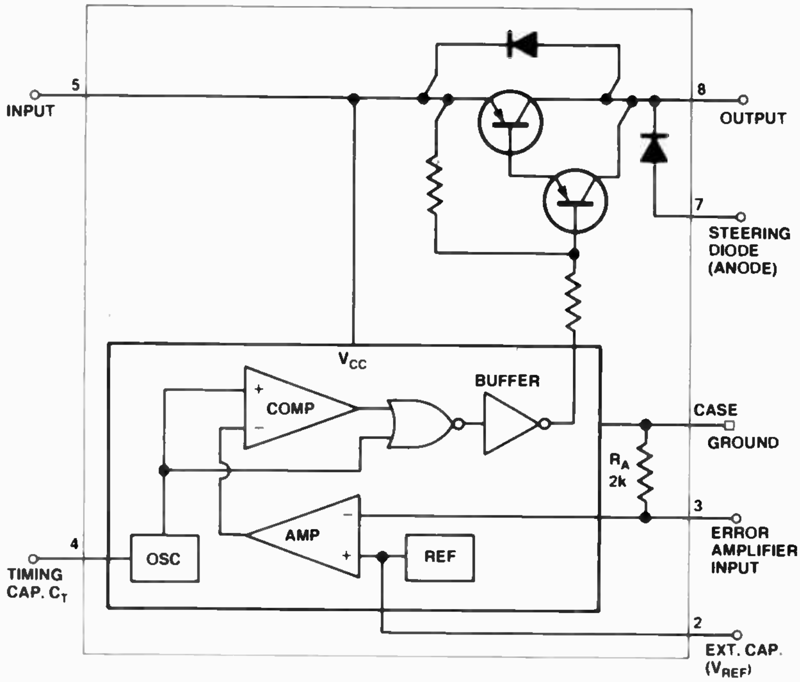
Barry Davis

THIS NEW regulator contains the power switching transistor, catch diode, and control circuitry to enable you to build an efficient switched mode power supply (SMPS). It has an adjustable output from 3 to 30 volts, a current capability of 5 amperes and is available in an 8-pin TO-3 package.

The features of the regulator are:

1. Output adjustable from 3 to 30 V.
2. 5 A output current
3. Step-down operation
4. High efficiency
5. Frequency of operation adjustable to 100 kHz
6. Maximum input voltage 35 V
7. Minimum input-output voltage differential = 5 V
8. TO-3 package.

The internal construction and the base connections of the integrated circuit are shown in Figure 1.



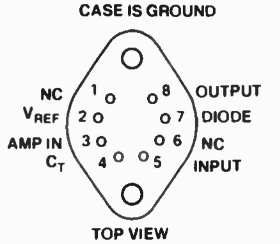


Figure 1. **Block diagram and base connections of the LH1605.**

Application details

The LH1605 is a step-down positive dc to dc switching converter with built-in control circuits, power transistor and catch diode. Few additional external components are required to provide an SMPS of 5 A capability. In the simplest design form, a timing capacitor (C_T) determines the operating frequency, an LC filter provides a low ripple dc output voltage and a single resistor sets the level of the output voltage. (R_s).

The value of R_s may be calculated from the formula:

$$R_s = 2\,k\Omega\, \frac{V_{out} - 2.5\,V}{2.5\,V}$$

A typical circuit is shown in Figure 2. R_s is made adjustable to allow the output voltage to be varied. Capacitor C1 is included to improve noise rejection by bypassing the reference (pin 2).

Operation

A basic step-down switching regulator circuit is shown in Figure 3.

Transistor Q is the switching element which has the 'on' and 'off' times controlled by a pulse width modulator (PWM). When *Q is on* power is drawn from the input and supplied to the load. The capacitor (C) charges and a voltage is developed across the inductor L. The diode (D) has a positive potential at its cathode, because Q is saturated, and is therefore held in reverse bias (off).

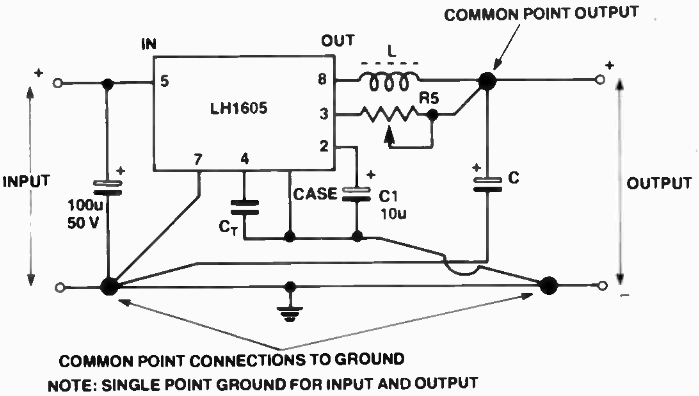


Figure 2. **Typical circuit applications.**

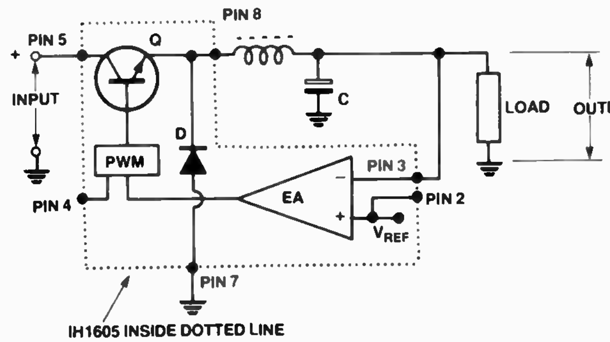


Figure 3. **Basic switched-mode power supply.**

When *Q is off,* the magnetic field stored in the inductor collapses. The polarity of the voltage across L reverses, forcing the diode into forward bias (on).

The energy stored in the inductor is released and the inductor current passes through the loop formed by the diode, inductor, and capacitor/load, thereby maintaining an output current.

The control circuit error amplifier (EA), samples the output voltage and automatically increases the 'on' time of Q if the output voltage decreases. Likewise it increases the 'off' time of Q if the output voltage increases.

This automatic action constantly compares the output voltage with a reference voltage and provides the necessary correction by varying the conduction time of Q.

With this type of circuit, the output can be calculated from the formula:

$$V_{out} = V_{in} \times \frac{t_{on}}{T}$$

where t_{on} = Q conduction time in seconds and T = period of one cycle of PWM (i.e. $t_{on} + t_{off}$).

Design procedure

The design of a complete voltage regulator with the LH1605 is relatively straightforward. However, prior to designing an operational circuit we must be aware of five circuit parameters. These are:

1. Maximum and minimum input voltage.
2. Required output voltage.
3. Maximum and minimum load current.
4. Maximum allowable ripple voltage.
5. Desired switching frequency.

Using the following parameters a design example can be given.

V_{in} (max) = 15 V
V_{in} = 10 V
V_{out} = 5 V
I_o (max) = 3 A
I_o (min) = 1 A
Output ripple Ωv_o = 20 mV
Operating frequency = 25 kHz

The values of the output L and C can be calculated, but firstly the 'off' time is required.

$$t_{off} = \frac{1 - \frac{V_{out}}{V_{in\,(max)}}}{f}$$

$$= \frac{1 - \frac{5}{15}}{25\text{ kHz}}$$

$$= 26.7 \text{ usecs}$$

The minimum equivalent frequency of the switching transistor at minimum input voltage is:

$$f(min) = \frac{1 - \frac{V_{out}}{V_{in\,(min)}}}{t_{off}}$$

$$= \frac{1 - \frac{5}{10}}{26.7 \text{ us}}$$

$$= 18.7 \text{ kHz}$$

The allowable peak to peak ripple current through L is:

$$\Delta i = 2 \times I_{o\,min}$$
$$= 2 \times 1\text{A}$$
$$= 2\text{A}$$

The inductance L can now be calculated.

$$L = \frac{V_{out} \times t_{off}}{\Delta i}$$

$$= \frac{5 \times 26.7 \text{ us}}{2\text{A}}$$

$$= 67 \text{ uH}$$

The value of minimum output capacitance is given by:

$$C = \frac{\Delta i}{8 \times f_{(min)} \times \Delta v_o}$$

$$= \frac{2\text{A}}{8 \times 18.7 \text{ kHz} \times 20 \text{ mV}}$$

$$= 688 \text{ uF (minimum)}$$

Because of the high value of ripple current, the capacitor must have a low equivalent series resistance (ESR):

$$ESR_{max} = \frac{\Delta v_o}{\Delta i}$$

$$= \frac{20 \times 10^{-3}}{2}$$

$$= 0.01 \text{ ohms}$$

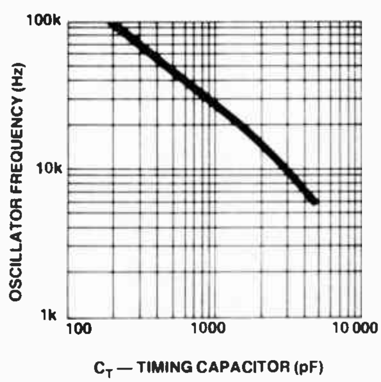


Figure 4. **Frequency vs. timing graph.**

Using the frequency/timing capacitor graph in Figure 4, a 0.001 uF (1 nF) capacitor used as the timing component will provide an operating frequency of 25 kHz. ▶

The output voltage of 5 V can now be programmed by calculating the value of Rs.

$$Rs = 2\,k\Omega\ \frac{V_{out} - 2.5}{2.5}$$
$$= 2\,k\Omega\ \frac{5 - 2.5}{2.5}$$
$$= 2\,k\Omega$$

The completed circuit is shown in Figure 5.

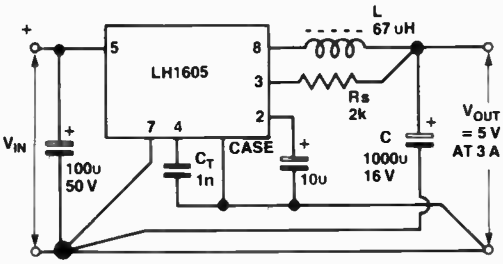


Figure 5. **Example circuit values.**

Choice of inductor core

The power handling capability of the inductor is reflected in the type of core material chosen. LI^2 must be calculated to determine the minimum useable inductor core *without* core saturation. L is the inductance previously calculated and I is the dc output current. It equals $Io_{(max)} + \Delta i$.

There are two popular core materials available —

1. molypermalloy
2. ferrite.

Ferrite potcores have the advantage of ease of winding and self-shielding against inherent magnetic fields. Molypermalloy cores are capable of higher flux density for a given core size.

Once the core is selected from the manufacturer's catalogue, the number of turns for the winding (N) can be calculated:

$$N = 1000 \sqrt{\frac{L}{L_{1000}}}$$

N = number of turns, L = inductance required. L_{1000} = inductance (mH/1000 turns) for the core, given by the manufacturer.

For example, in the design exercise, the desired value of L was 67 uH:

$$\text{Peak } I^2 = (3\,A + 2\,A)^2 = 25\,A$$

Therefore:

$$LI^2 = 67\,uH \times 25\,A = 1.68 \text{ millijoules}$$

The permalloy core chosen has a nominal inductance of 120 mH per 1000 turns. Therefore:

$$N = 1000 \sqrt{\frac{67\,uH}{120\,mH}} = 24 \text{ turns}$$

Heatsinking

There will be significant self-heating due to internal power dissipation, even at moderate output power. The junction temperature must be kept as low as possible under operating conditions (maximum temperature 150°C). Good heatsinking is essential and a mica washer complete with 'thermal' paste is recommended, along with a TO-3 styled heatsink.

Short circuit protection

The LH1605 will be *permanently damaged* if under a short circuit condition for longer than 10 milliseconds, and it is therefore desirable to add current limit protection to the circuit. A current limit circuit is shown in Figure 6.

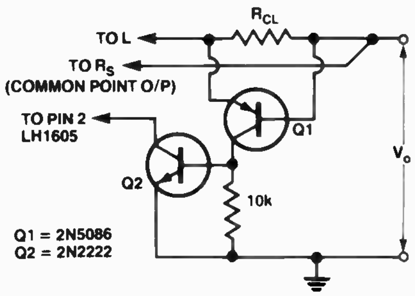


Figure 6. **Short circuit protection.**

The value of R_{CL} is determined by the current limit you decide to set. It can be calculated from the formula:

$$R_{CL} = \frac{660\,mV}{I_{sc}}$$

where I_{sc} = short circuit current.

The voltage developed across R_{CL} by the short circuit current (600 mV) will turn transistor Q1 on. The increase in I_c of Q1 will develop a voltage across the 10k resistor, making the base of Q2 more positive with respect to the emitter. Q2 will turn on, pulling the reference of the control circuit (pin 2) down towards ground. The error amplifier suddenly sees too high an output voltage. This shuts the series pass transistor off in the LH1605 by reducing the switching time, thereby reducing the output voltage to the level of the reference.

A complete power supply with current limiting at 3.25 A is shown in Figure 7.

Evaluation kit

A *complete* evaluation kit including the printed circuit board, inductor, and specification sheet is available from:

National Semiconductor
Cnr. Stud Road & Mountain Highway
Bayswater
Vic. 3153.

The evaluation kit, SR1605, is fully assembled and ready for immediate use. It is designed to accept any unregulated input voltage from 10 to 25 V and provide a 5 V 3A output.

Acknowledgement

This article was made possible by the courtesy of National Semiconductor. Data was taken from their publication: LH1605 5 Amp High Efficiency Switching Regulator.

Further information on the operation of switched mode power supply circuits can be obtained from the book 'Understanding dc Power Supplies', by Barry Davis; publisher: Prentice Hall. ●

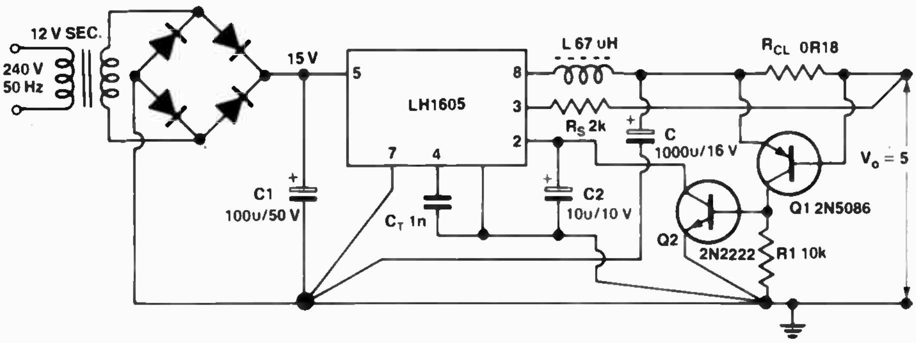


Figure 7. **5 V SMPS with short circuit protection.**

Which battery to use?

The advent of microcircuits leading to miniaturisation of equipment and to decreased power demands has resulted in a resurgence of battey operated devices. Advances in technology have led to some remarkable developments in batteries today. But there you have a problem. Which battery to use?

Philip Clark

MANY PEOPLE have a tendency to replace conventional 'dry' cells with the rechargeable nickel-cadmium type simply to avoid replacement costs, but this practice may not be cost effective and can even lead to reduced equipment performance and greater cost. In order to make some form of meaningful comparison between types, it was necessary to select one common size of cell and look at the performance under similar load conditions. Also, because the performance of carbon-zinc cells is more dependent on usage than alkaline or NiCad cells, a definite statement as to which is best cannot readily be made.

The purpose of this article is to try to present a guide to the selection of cell type best suited for an application, bearing in mind equipment performance, duty cycle, current demand, weight and cost. The information presented here has been obtained from the data sheets of many manufacturers and should not be regarded as typical of any particular make. Because of the variation in performance of carbon-zinc cells with differing loads, the presentation has been optimised to give a reasonable overall guide to performance without being too optimistic.

Cells tested

The basic cells compared are the nickel-cadmium rechargeable type, the ordinary carbon-zinc dry cell, both normal and heavy duty (leclanche type) and the manganese dioxide-alkaline type. The size of the cells selected for comparison are the 'AA' or UM-3, SAA designation R6. In the case of the NiCad this is a 450 mA hour capacity cell. The load characteristic selected was a current drain of 30 mA for four hours per day. This is typical of much portable equipment such as walkie-talkies, portable radios, calculators, etc. This was also the load which was easiest to compare on a range of data for various cells.

Result of comparisons

Generally, the comparison here shows alkaline cells to have a performance about twice that of ordinary carbon-zinc cells, however they can be many times better, depending on usage. Alkaline cells have higher efficiency when used for continuous or heavy load (high current) applications where the conventional carbon-zinc cell is less effective. The rechargeable NiCad is good for heavy current applications, provided the correct supply voltage can be achieved. Unfortunately, carbon-zinc and alkaline cells are not directly interchangeable with NiCad cells due to ▶

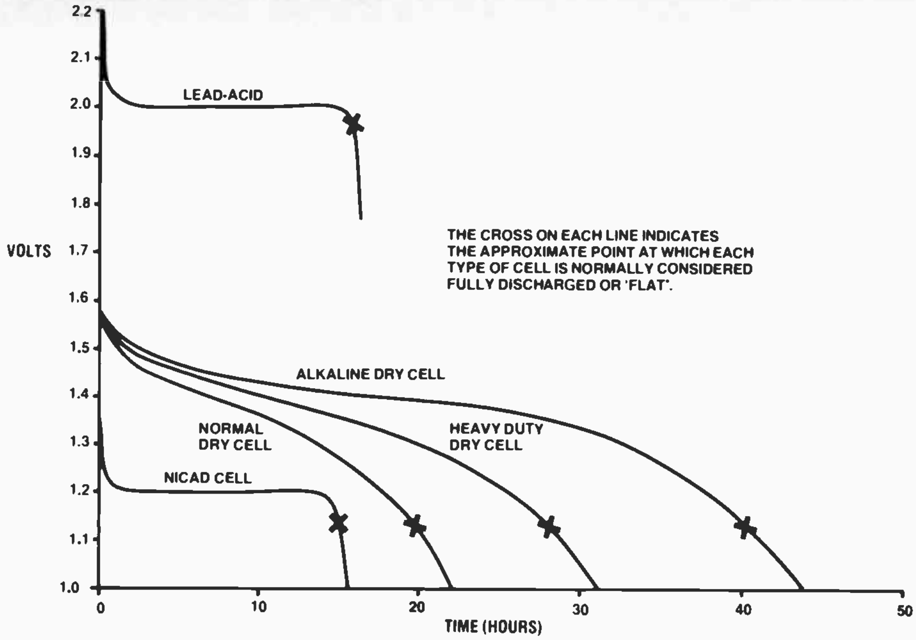


Figure 1. **The discharge performance of various cells. The discharge characteristics of a lead-acid cell are shown for comparison. Note that terminal voltage cannot be used to indicate the state of charge of NiCad cells.**

differing terminal voltages. This is caused by the different types of materials and construction used for these cells. The result is that while carbon-zinc and alkaline cells have a terminal voltage of 1.5 V, the NiCad cell has a voltage of only 1.2 V.

Figure 1 shows the performance of the various cells and that of a similar capacity lead-acid cell for comparison. It is important to note that the terminal voltage cannot be used to indicate the state of charge of the NiCad cell but can be used as an approximate indicator for 'dry' cells. NiCad cells are about 25% to 30% heavier than 'dry' cells and because more of them may be required for the same voltage, this could mean a substantial weight penalty in portable equipment.

Batteries at work

The place where these various cells must work is in the equipment, and this is where many factors become important. The points to be considered are: *duty cycle* — is the load to be continuous or intermittent high current, or is it to be low current, continuous or intermittent? *Operating environment*— will the power be required at extremes of temperature? What is the design voltage of the equipment and can sufficient cells be accommodated to provide this? *Replacement or recharging* — in a particular situation one option may be preferable to the other, and which option is cost effective? *Operational life*— how long will the selected cell operate before recharge or replacement is necessary? *Shelf or storage life*— how good is a particular cell after a period of no use?

Duty cycle

This will have a major effect on the performance of a cell in any situation. Conventional carbon-zinc cells perform best at a relatively light load when operated intermittently. This allows a degree of recovery between periods of use. The service life of alkaline cells is relatively constant regardless of whether use is continuous or intermittent. This type of cell then shows its advantage mainly when continuous use is required. It can have a service life of three to ten times that of carbon-zinc cells in ideal circumstances.

Another advantage of the alkaline cell is its ability to supply considerably higher currents than the carbon-zinc type. In fact, the current available from alkaline cells can approach that from NiCad rechargeable cells in some circumstances. For high current loads, intermittent or continuous, the NiCad cell may be preferred, either because other cell types cannot supply the required current or because discharge is so rapid that continual replacement would be necessary. Substitution of carbon-zinc or alkaline cells with NiCads should only be undertaken after consideration of all the factors involved, including operating environment, equipment voltage requirements and storage life.

Operating environment

Carbon-zinc cells deteriorate quickly at temperatures above about 50°C and become rapidly unable to deliver useful current below −18°C. Alkaline cells show better operating characteristics at extremes of temperature. Although it is difficult to determine the upper temperature limit of these cells, it is considerably better than carbon-zinc cells. Alkaline cells perform reasonably well down to temperatures of −40°C.

NiCad rechargeable cells have an operating temperature range of about −20°C to +45°C but should not be exposed to temperatures below 0°C while charging. Generally, their operating temperatures are about the same as for carbon-zinc cells. There may be some temperature rise in NiCad cells during charging or heavy discharge, and this factor should be considered if these cells are used as a replacement for 'dry' cells in sensitive or critical equipment.

Design voltage of equipment

This is an important factor when consideration is being given to replacing carbon-zinc or alkaline cells with rechargeable NiCad types or replacing NiCads with 'dry' cells. In equipment such as portable transceivers, satisfactory operation depends greatly on the available supply voltage. Some of this equipment is designed to operate from NiCad cells while other is intended to use 'dry' cells. Most such equipment has a specified operating range of voltages and attempted use outside of this range will result in severely degraded performance, no operation at all, or possible damage to the equipment. Typical ranges for nominal 12 V equipment are 11 V to 14 V or 10 V to 15 V. There is a temptation to replace carbon-zinc or alkaline cells with an equal number of rechargeable NiCad cells but because of the difference in terminal voltage (1.5 V as compared to 1.2 V), a fully charged NiCad battery may not meet the minimum voltage requirements of equipment.

Figure 2 shows the performance of various battery types in a piece of 12 V rated equipment such as a portable transceiver. Note that if provision is only made for eight cells, then replacement with lower voltage types can result in insufficient supply for the correct operation of the equipment. Often, the mere fact that the equipment operates at all under these conditions is more a tribute to the designer than the performance of the power source! On the other hand, replacement of NiCad cells with 'dry' cells could allow the equipment voltage specification to be exceeded.

Replacement or recharging

Which is best here will depend very much on the user requirement. For example, a transceiver used by emergency

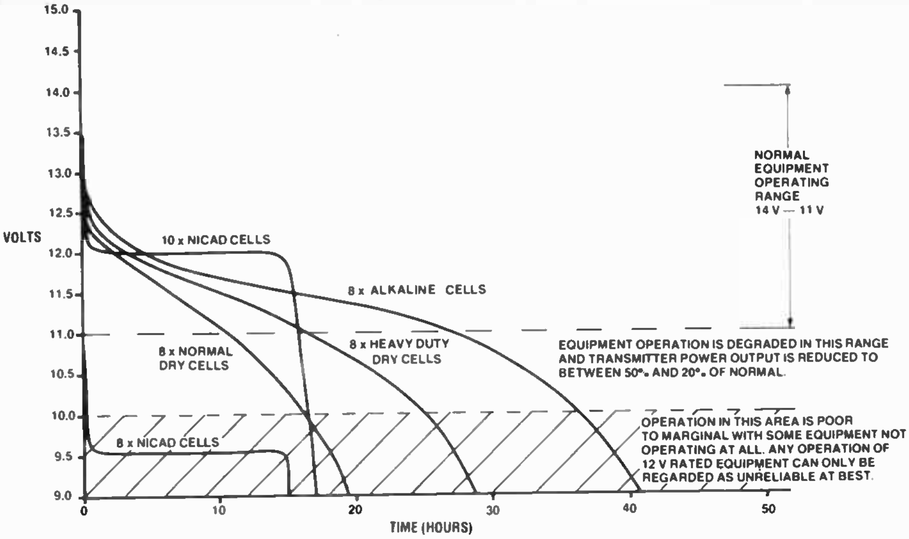


Figure 2. **The performance of various battery types in a 12 V operated handheld transceiver. Note that you cannot expect to replace eight dry cells with eight NiCads.**

services might be more quickly restored to service by replacement of batteries than by recharging. For personal use, where failure due to battery discharge is not so critical, recharging may be acceptable. If the equipment is in heavy or continuous use then recharging may be a viable alternative to replacement.

The cost of any option will depend on how often replacement is required, the higher initial cost of rechargeable cells, (and you may need more of them), the cost of a charger and the cost of an additional battery pack if operation is needed while recharging is in progress. Another important factor, especially in an emergency environment, is the availability of power for recharging. If such a source is not readily accessible, a user may be ill-advised to use rechargeable batteries at all.

Operational life

The effective operational life of a battery may become very important if access to replacement or recharged cells is difficult. As can be seen from Figure 2, the NiCad cells under this load will supply power for about 15 hours of operation. This is conditional on the battery being fully charged and in good condition. Ordinary carbon-zinc cells will provide about the same service life and heavy duty ones about twice this. Alkaline cells can be expected to have a greater operating life, about three times or better, than either NiCads or carbon-zinc as compared here.

There is a further factor which should be considered where the use of NiCad cells is contemplated. It has become recognised that NiCad cells tend to develop a 'memory' of their usage pattern. What appears to happen is that if a cell is used to say only 50% of capacity and then recharged, after a few cycles of this pattern the cell then becomes only capable of delivering 50% of rated capacity before going 'flat'. This condition can be reversed by correctly cycling the cell through several discharge and charge cycles, but unless this condition is recognised as developing, it may seriously degrade the operational life of the equipment in which it is used.

Shelf life or self discharge

All cells will discharge by themselves when not in use, to a greater or lesser degree. This self discharge will determine the length of time for which a cell can be left unused and still be able to deliver a reasonable proportion of its original charge. The major factors which influence the rate of self discharge are storage temperature, amount of charge at storage, and the condition of the cell.

The definition of shelf life is somewhat variable, but for carbon-zinc and alkaline cells appears to be the time taken to decrease to 90% of initial capacity. Accepting this definition then gives the following approximate storage lifetimes for cells in good condition.

Carbon-zinc cell — 8 to 9 months
Alkaline cell — Over 2 years.
NiCad cell — 3 days to 4 weeks.

These storage times are based on a constant temperature of about 20°C to 25°C. Storage life may be improved by storing the cells at 5-10°C. Generally, higher temperatures cause more rapid degradation. Storage life is also shown in graphical form in Figure 3. An approximation for a lead-acid car battery type cell is shown for comparison. NiCad cells appear to show up poorly in this regard and some manufacturers now claim to have substantially improved this characteristic. The self discharge of NiCad cells depends on the type of cell, whether it is intended for high or normal current discharge. The condition of the cell is also important, whether it has been cycled correctly, the age (number of cycles), the environment in which it has been used and the state of charge at storage.

Consideration of the information provided here may be able to assist you to make a better informed decision as to the best power source for your battery operated equipment. It is not practical to cover all the eventualities and applications in a short article, but at least this should provide some guide to the cost effectiveness and practicability of the battery that you select. Most manufacturers will provide design and engineering data on request should you need to make a more detailed analysis of your particular needs. ●

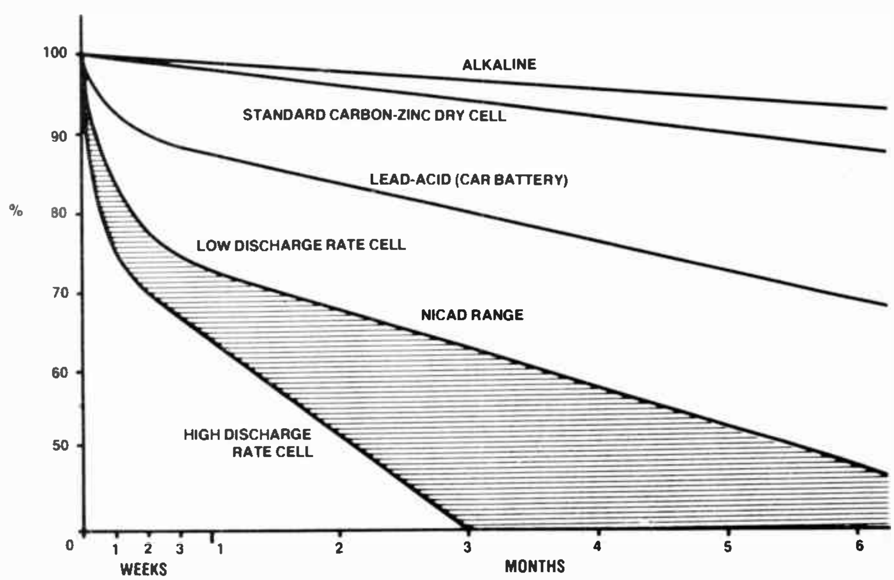
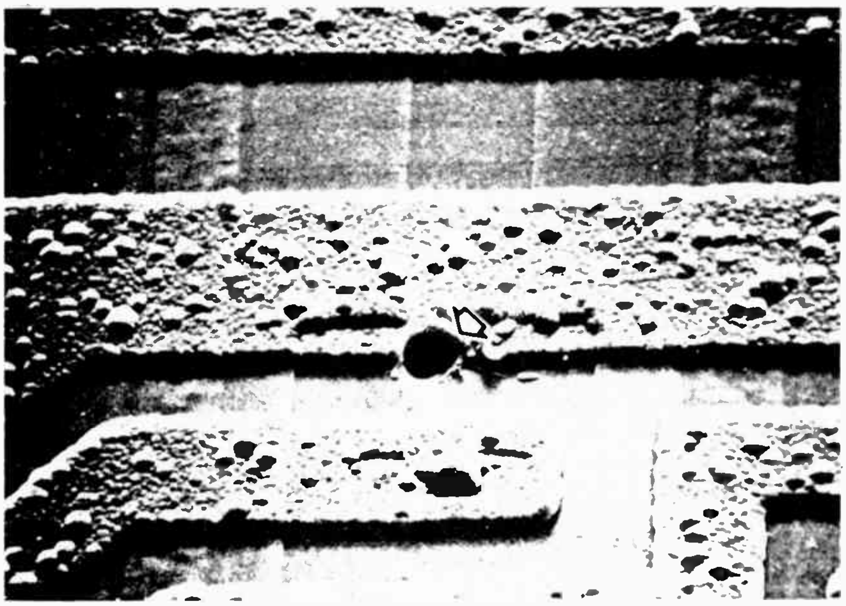


Figure 3. **All secondary cells self-discharge, which gives them a certain 'shelf life'. Here, the self-discharge characteristics of the various cells are shown. A lead-acid cell is included for comparison.**

Electrostatic discharge —nemesis of electronic systems

Electrostatic hazards to electronic equipment exist during assembly and testing as well as subsequent shipment and then handling by the user. To protect these systems, it is necessary to understand the nature of electrostatic discharge (ESD), how it is generated, and how it is transmitted to electrostatic discharge sensitive components.

D.E.Frank



Close up view (x1000) of electrostatic damage in an op-amp.

What is ESD?

OUR MOST common conception of electrostatic discharge is the miniature lightning bolt or shock we receive in periods of low humidity when we walk across a carpet or slide across vinyl seatcovers and then discharge to a door knob or door handle. In essence, however, most static electricity is subliminal or occurs at values well below our perception level of 1500 to 2000 volts.

Static electricity is generated whenever two substances are rubbed together, separated, or flow relative to one another (such as a gas or liquid over a solid). This static electricity (electrical charge at rest) is stored on non-conductive materials and tends to remain in the localised area of contact awaiting the opportunity to discharge to the first available ground source. In the case of conductive materials, the charge is rapidly distributed over the entire surface and the surface of other conductors which come in contact.

The Triboelectric Series (Table 1) shows the charge relationships of many materials. Note that cotton is identified as a reference material, being at mid-point of Table 1. It tends to absorb moisture, thereby rendering it somewhat conductive. However, when cotton is rubbed against another material, it has the ability to produce a static charge.

Materials listed above cotton tend to assume a positive charge by giving off electrons in a friction separation situation, while those listed below cotton become negatively charged by acquiring electrons. When any two materials experience separation or rubbing, the material listed highest on the table will become positively charged, and the material listed lower will accept the negative charge.

For the sake of simplicity, let us define the cause of common static as the flow of materials and people within an environment. Materials include all components, packaging, and other raw materials which make up our finished product. People carry and generate charges, and it all takes place within a defined environment made up of facilities and equipment. The environment is not limited to a plant, but can be defined as a package, or many plants as in a transport situation.

Static, as it manifests itself in our environment, is actually a symptom. If we can impose control on the elements which create static as an end result, we can control the generation of a myriad of problems caused by static.

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It is the author's intent to present an understanding of ESD, to explain how various materials and situations generate ESD, and to discuss ESD sensitive devices in depth. In the discussion of ESD sensitive devices, an effort has been made to present a thorough "physics of failure" analysis to provide insight into the design and structure of ESD sensitive devices as well as the failure modes and effects. Those individuals not desiring an in-depth technical treatise may review Tables 2 and 3, and also the discussion in paragraphs entitled "Detecting ESD Damage" and "Protecting Your Equipment".

ESDS devices

Typical ESD voltages are shown in Table 2, and the ranges of susceptibility of ESDS devices are shown in Table 3. Parts of devices can be destroyed (hard failures) or simply degraded or made intermittent (soft or upset failures) due to exposure to electrostatic discharge. Parts are susceptible to damage when an ESD occurs across their terminals. ESDS parts can be destroyed by an ESD where one pin is connected to a high voltage source and other pins are ungrounded. In other words, a hard ground connection is not required to destroy an ESDS part.

MOS large scale integration devices in hermetic packages with non-conductive lids could be damaged by spraying the lids with canned coolant, despite there being no ground path connected to the part.

MATERIALS	
AIR	Increasingly Positive
HUMAN HANDS	
ASBESTOS	
RABBIT FUR	
GLASS	
MICA	
HUMAN HAIR	
NYLON	
WOOL	
FUR	
LEAD	
SILK	
ALUMINUM	
PAPER	
COTTON	
STEEL	Increasingly Negative
WOOD	
AMBER	
SEALING WAX	
HARD RUBBER	
NICKEL COPPER	
BRASS SILVER	
GOLD PLATINUM	
SULFUR	
ACETATE RAYON	
POLYESTER	
CELLULOID	
ORLON	
SARAN	
POLYURETHANE	
POLYETHYLENE	
POLYPROPYLENE	
PVC (vinyl)	
KEL-F (CTFE)	
SILICON	
TEFLON	

Table 1. Triboelectric series

Means of Static Generation	Electrostatic Voltages 10 to 20% Relative Humidity	65 to 90% Relative Humidity
Walking across carpet	35 000	1 500
Walking over vinyl floor	12 000	250
Worker at bench	6 000	100
Vinyl envelopes for work instructions	7 000	600
Common poly bag picked up from bench	20 000	1 200
Work chair padded with polyurethane foam	18 000	1 500

Ref. U.S. Department of Defense DOD-HDBK-263 (tables III and IV) 2 May. 1980.

Table 2. Typical electrostatic voltages

ESDS parts installed in assemblies normally have their leads connected to a sufficient mass of conductive material, such as printed circuit board (pcb) runs and wiring, which may provide the required ground to result in damage from an ESD. In such cases, however, the voltages required are normally higher than those needed when one or more pins or the part case is grounded.

Assemblies and equipment containing ESDS parts are often as sensitive as the most sensitive ESDS part which they contain. Incorporation of protective circuitry in these assemblies and equipment can provide varying degrees of protection from ESD applied to their terminals. Such assemblies and equipment, however, can still be vulnerable from induced ESD caused by strong electrostatic fields or by contact of pcb electrical connections or paths with a charged object.

Intermittent or upset failures can occur on certain types of parts, such as LSI memories and chips, either prior to or after lidding and sealing. Such failures can also occur when equipment is in operation, characterised by a loss of information or temporary distortion of its functions. No apparent hardware damage occurs and proper operation resumes automatically after the ESD exposure or, in the case of some digital equipment, after re-entry of the information by resequencing the equipment.

Upset can also be the result of an ESD spark in the vicinity of the equipment. The electromagnetic pulse generated by the spark causes erroneous signals to be picked up by the equipment circuitry. Upset can also occur by the capacitive or inductive coupling of an ESD pulse or by the direct discharge of an ESD through a signal path providing an erroneous signal.

While upset failures occur when the equipment is operating, catastrophic failures can occur any time. Catastrophic ESD failures can be the result of electrical overstress of electronic parts caused by an ESD, such as: a discharge from a person or object, an electrostatic field, or a high voltage spark discharge (see Figure 1).

Some catastrophic failures may not occur until some time after exposure to an ESD, as in the case of marginally damaged ESDS parts, which require operating stress and time to cause further degradation and ultimate catastrophic failure. Only certain part types seem to be susceptible to this latent failure process.

There are some types of catastrophic ESD failures which could be mistaken for upset failures. For example, an ESD could result in aluminium shorting through a SiO_2 dielectric layer. Subsequent high currents flowing through the short, however, could vaporise the aluminium and open the short. This failure may be confused with upset failure if it occurs during equipment operation, but the damage due to the ESD would cause a latent defect that will probably reduce the operating life of the part.

Parts that are very susceptible to ESD upset are any within logic families that require small energies to switch states or small changes of voltage in high impedance lines. Examples of families that are sensitive would be NMOS, PMOS, CMOS, and low power TTL. Linear circuits with high impedance and high gain inputs would also be highly susceptible, along with RF amplifiers and other RF parts at the equipment level; however, design for RFI immunity can protect these parts from damage due to ESD high voltage spark discharge.

To protect parts sensitive to ESD high spark discharge at the equipment level requires: good radio frequency interference (RFI)/electromagnetic compatibility (EMC) design, buffering of busses, proper termination of busses, shielding of buss conductors, and the avoidance of penetrations of the equipment enclosure that lead to sensitive parts. ▶

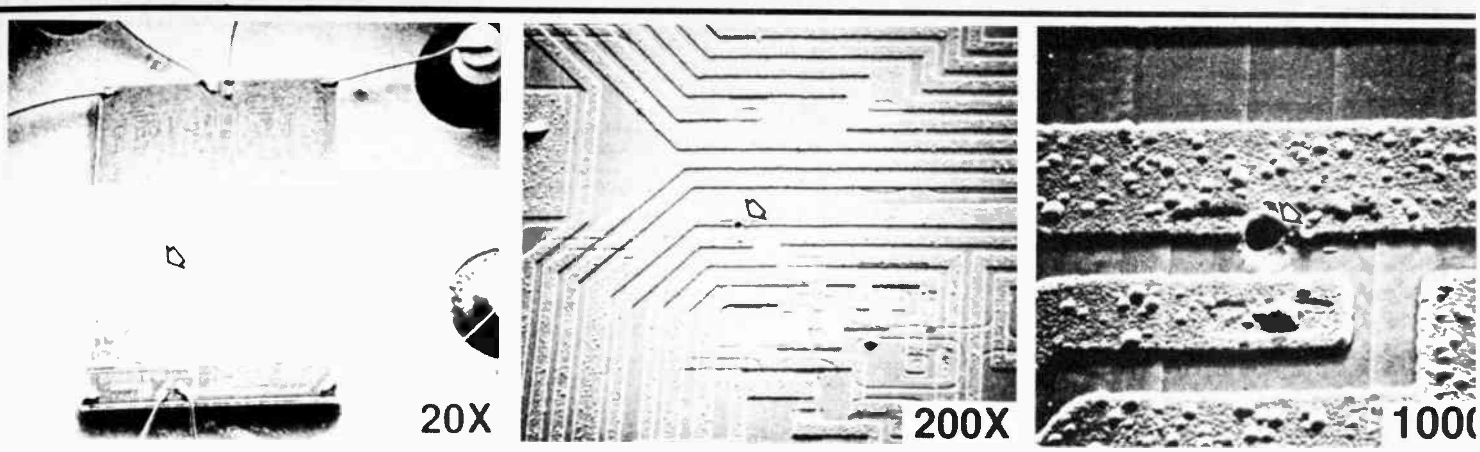


Figure 1. **Static discharge damage in an op-amp integrated circuit (arrows show location of damage).**

Failure mechanisms

Typical ESD failure mechanisms are divided into the two following categories. Those in the first category, thermal secondary breakdown, metallisation melt and bulk breakdown, are power dependent. Those in the second category, dielectric breakdown, gaseous arc discharge, and surface breakdown, are all voltage dependent. All of the above are applicable to microelectronic and semiconductor devices. Metallisation melt and gaseous arc discharge are evident in film resistors, and bulk breakdown is typical of piezoelectric crystals.

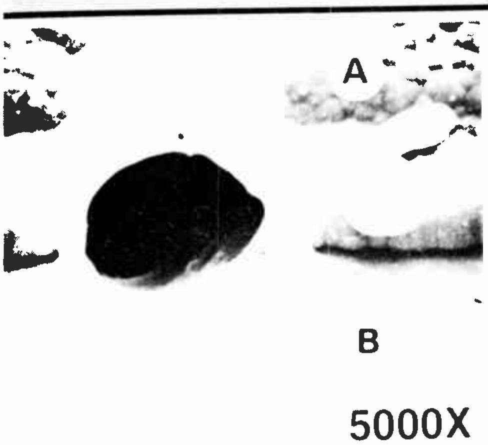
Besides these catastrophic failure mechanisms, un-encapsulated chips and LSI MOS integrated circuits have exhibited temporary failure due to gaseous arc discharge from positive charges deposited on the chip as a by-product of gaseous arc discharges within the package between the lid and the substrate.

Thermal secondary breakdown is also known as avalanche degradation. Since thermal time constraints of semiconductor materials are generally large compared with transient times associated with ESD pulse, there is little diffusion of heat from the areas of power dissipation, and large temperature gradients can form in the parts. Localised junction temperatures can approach material melt temperatures, usually resulting in development of hot spots and subsequent junction shorts due to melting.

For junction melting to occur in bipolar (P-N) junctions, sufficient power must be dissipated in the junction. In the reverse bias condition, most of the applied power is absorbed in the immediate junction area with minimal power loss in the body of the part. In the forward bias condition, the junction in-

CLASS 1: SENSITIVITY RANGE 0 TO <1000 VOLTS
Metal Oxide Semiconductor (MOS) devices including C (Complementary), D (Double-Diffused), N (N-Channel), P (P-Channel), V (V-Groove) and other MOS technology without protective circuitry, or protective circuitry having Class 1 sensitivity
Surface Acoustic Wave (SAW) devices
Operational Amplifiers (OP AMP) with unprotected MOS capacitors
Junction Field Effect Transistors (JFETs) (Ref.: Similarity to MIL-STD-701 Junction field effect, transistors and junction field effect transistors, dual unitized)
Silicon Controlled Rectifiers (SCRs) with I_O < 0.175 amperes at 100° Celsius (°C) ambient temperature (Ref.: Similarity to MIL-STD-701 Thyristors (silicon controlled rectifiers))
Precision Voltage Regulator Microcircuits: Line or Load Voltage Regulation <0.5 percent
Microwave and Ultra-High Frequency Semiconductors and Microcircuits: Frequency >1 gigahertz
Thin Film Resistors (Type RN) with tolerance of <0.1 percent; power >0.05 watt
Thin Film Resistors (Type RN) with tolerance of >0.1 percent, power ≤0.05 watt
Large Scale Integrated (LSI) Microcircuits including microprocessors and memories without protective circuitry, or protective circuitry having Class 1 sensitivity (Note LSI devices usually have two to three layers of circuitry with metallization crossovers and small geometry active elements.)
Hybrids utilizing Class 1 parts
CLASS 2: SENSITIVITY RANGE >1000 TO ≤4000 VOLTS
MOS devices or devices containing MOS constituents including C, D, N, P, V, or other MOS technology with protective circuitry having Class 2 sensitivity
Schottky diodes (Ref.: Similarity to MIL-STD-701 Silicon switching diodes (listed in order of increasing trr))
Precision Resistor Networks (Type R2)
High Speed Emitter Coupled Logic (ECL) Microcircuits with propagation delay ≤1 nanosecond
Transistor-Transistor Logic (TTL) Microcircuits (Schottky, low power, high speed, and standard)
OP AMPs with MOS capacitors with protective circuitry having Class 2 sensitivity
LSI with input protection having Class 2 sensitivity
Hybrids utilizing Class 2 parts
CLASS 3: SENSITIVITY RANGE >4000 TO ≤15,000 VOLTS
Lower Power Chopper Resistors (Ref Similarity to MIL-STD-701 Silicon Low Power Chopper Transistors)
Resistor Chips
Small Signal Diodes with power ≤1 watt excluding Zeners (Ref. Similarity to MIL-STD-701 Silicon Switching Diodes (listed in order of increasing trr))
General Purpose Silicon Rectifier Diodes and Fast Recovery Diodes (Ref Similarity to MIL-STD-701 Silicon Axial Lead Power Rectifiers, Silicon Power Diodes (listed in order of maximum DC output current), Fast Recovery Diodes (listed in order of trr))
Low Power Silicon Transistors with power ≤5 watts at 25°C (Ref Similarity to MIL-STD-701 Silicon Switching Diodes (listed in order of increasing trr), Thyristors (bi-directional triodes), Silicon PNP Low-Power Transistors (Pc ≤5 watts @T_A = 25°C), Silicon RF Transistors)
All other Microcircuits not included in Class 1 or Class 2
Piezoelectric Crystals
Hybrids utilizing Class 3 parts

Table 3. List of ESDS devices by part type



Detailed view of the 6-micron (0.0002") diameter hole created in aluminium metallisation (A) and silicon dioxide substrate (B) by static discharge.

hibits lower resistance. Even though a greater current flows, a greater percentage of the power is dissipated in the body of the part. Thus, more power is generally required for junction failure in the forward bias condition.

For most transistors, the emitter-base junction degrades with lower current values than the collector-base junction. This is because the emitter-base junction normally has smaller dimensions than any of the other junctions in the circuit. For reversed polarity signals, only a very small microampere current flows until the voltage exceeds the breakdown voltage of the junction. At breakdown, the current increases and results in junction heating, due to the nucleation of hot spots and current concentrations. At the point of second breakdown, the current increases rapidly due to a decrease in resistivity and a melt channel forms that destroys the junction. This junction failure mode is a power-dependent process.

Metallisation melt failures can occur when ESD transients increase part temperature sufficiently to melt metal or fuse-bond wires. Theoretical models exist which allow computation of currents that can cause failure for various materials as a function of area and current duration. Such models are based on the assumption of uniform area of the interconnection material. In practice, it is difficult to maintain a uniform area, and the resultant non-uniform area can result in localised current crowding and subsequent hot spots in the metallisation. This type of failure could occur where the metal strips have reduced cross-sections as they cross oxide steps. Normally due to shunting of the currents by the junction, this failure requires a larger power level at higher frequencies than is required for junction damage at lower frequencies. Below 200 to 500 MHz, the junction capacitance still presents a high impedance to currents, thereby shunting them around the junction.

Bulk breakdown results from changes in junction parameters due to high local temperatures within the junction area. Such high temperatures result in metallisation alloying or impurity diffusion, resulting in drastic changes in junction parameters. The usual result is the formation of a resistance path across the junction. This effect is usually preceded by thermal secondary breakdown.

Dielectric breakdown occurs when a potential difference is applied across a dielectric region in excess of the region's inherent breakdown characteristics, and a puncture of the dielectric occurs. This form of failure is due to voltage rather than power and could result in either total or limited degradation of the part, depending on the pulse energy. For example, the part may heal from a voltage puncture if the energy in the pulse is insufficient to cause fusing of the electrode material in the puncture. It will, however, usually exhibit lower breakdown voltage or increased leakage current after such an event, but it will not exhibit catastrophic part failure.

This type of failure could result in a latent defect resulting in catastrophic failure with continued use. The breakdown voltage of an insulating layer is a function of the pulse rise time, since time is required for avalanching of the insulating material.

Gaseous arc discharge occurs in parts with closely spaced, unpassivated, thin electrodes. Gaseous arc discharge can cause degraded performance. The arc discharge condition causes vaporisation and metal movement, which is generally away from the space between the electrodes. The melting and fusing do not move the thin metal into the interelectrode regions. In melting and fusing, the metal flows together and flows or opens along the electrode lines. There can be fine metal globules in the gap region, but not in sufficient numbers to cause bridging. Shorting is not considered a major problem with passivated thin metal electrodes.

On a SAW band pass filter device with thin metal of approximately 0.4 micron and 3.0 microns (1 micron = 1 x 10^{-6} metres) electrode spacing, operational degradation was experienced from ESD. When employing thicker metallisation such as 1.35 microns, this gaseous arc discharge in an arc gap at typically 50 microns can be used for protection to dissipate incoming high voltage spikes.

For LSI and memory ICs with passivation/active junction interfaces susceptible to inversion, gaseous arc discharge from inside the package can cause positive ions to be deposited on the chip and cause failure from surface inversion. This has been reported to occur especially on parts with nonconducting lids. A special case of this is UV-EPROMs with quartz lids, where failures can be annealed by neutralising the positive charge with ultraviolet light through the quartz lid.

Surface breakdown occurs at perpendicular junctions, and is explained as a localised avalanche multiplication process caused by narrowing of the junction space charge layer at the surface. Since surface breakdown depends on numerous variables, such as geometry, doping level, lattice discontinuities, and unclean gradients, the transient power which can be dissipated during surface breakdown is generally unpredictable.

The destruction mechanism of surface breakdown results in a high leakage path around the junction, thus nullifying the junction action. This effect, as well as most voltage sensitive effects like dielectric breakdown, is dependent upon the rise time of the pulse and usually occurs when the voltage threshold for surface breakdown is exceeded before thermal failure can occur.

Another mode of surface failure is the occurrence of an arc around the insulating material. This failure is similar to metallisation gaseous discharge except that discharge is between metallisation and semiconductor.

Specific effects on circuits

Now, having identified the causes of ESD and the major types of failure mechanisms, it is important to assess how these failures manifest themselves in systems. Typical devices and their degradation thresholds are summarised in Table 4. A more detailed analysis of the physical mechanisms follows.

MOS structures are a conductor and semiconductor substrate separated by a thin dielectric. This family includes MOS field effect transistors (FETs), MOS ICs, bipolar, hybrid, linear and digital ICs and MOS capacitors. Or more basically, the family includes any dual dielectric system or semiconductor with metallisation crossovers. The newer devices in this area — the VMOS (vertical groove MOS), the HMOS (high density MOS), the HEX MOS, and some of the prototype GaAs MESFETS (gallium arsenide metal semiconductor field effect transistor) — approach 1 micron or less compared with today's chip geometries of 4 to 5 microns (see Figure 2). Needless to say, as these ▶

DEVICE	TEST RESULTS	
	THRESH-OLD,[1] VOLTS	DEGRADATION CRITERIA[2]
Diodes		50% drop in V_R at I_R = 5 μA
1N459	> 3000	
1N916	3000	
TI551	450	
1N4151	> 3000	
Zener Diodes		50% drop in V_R at I_R = 5 μA
LVA356	> 3000	
Transistors		50% drop in $V_{(BR)CBO}$ at I_R = 5 μA
2N2222	1000	
2N2369A	460	
2N2432A	620	
2N2540	1450	
2N2907	1200	
2N3117	1000	
2N3570	380	
2N4251	460	
2N4872	1200	
2N5154	> 3000	
Junction Field-Effect Transistors		50% drop in $V_{(BR)GSS}$ at I_G = 5 μA
2N2608	320	
2N3112	530	
2N3971	160	
2N4118A	140	
Metal-Oxide Semiconductor Transistors		I_G > 5 μA at V_{GS} = 22 V
GI MEM 520c (chip)	58	
Complementary Metal-Oxide Semiconductor Integrated Circuits		> 0.5 μA input at 10 V or > 10% decrease in output voltage across 100-KΩ load
RCA CD4001	250	
Silicon-Controlled Rectifiers		50% increase in I_{CGO}
2N886A	680	
2N3030	1000	

NOTES:
1. Reverse-breakdown polarity.
2. Where V_R is the reverse voltage, I_R the reverse current, $V_{(BR)CBO}$ the collector/base breakdown voltage, I_B the base current, $V_{(BR)GSS}$ the gate/source breakdown, I_G the gate current, V_{GS} the gate/source voltage, and I_{CGO} the gate leakage current.

Table 4. Typical device degradation threshold

smaller geometries blend with higher purity processing, the device susceptibilities will rise and ESD transients of 20 volts will become lethal.

Differences in susceptibilities of these MOS technologies are dependent upon the gate dielectric strength and the oxide thickness. In the past, gate dielectric thickness has typically been around 0.11 micron with dielectric strengths ranging around 1 x 10^6 to 1 x 10^7 V/cm, with breakdown between 80 and 120 volts. Researchers today, however, are creating functional devices with dielectric thicknesses in the 0.06 to 0.08 micron range and breakdowns at 20 and 25 volts.

Many monolithic ICs have metallisation runs which cross over active semiconductor regions with field oxide between them serving as the insulator. These are called parasitic MOS transistors. Normally, these break down around 100 volts due to field intensification at the corners of the metallisation and weak dielectric strength of the oxide barrier. Breakdown of the oxide insulator is permanent, as opposed to breakdown of a semiconductor, which is reversible.

If very short-term overvoltages occur, a subsequent breakdown or avalanche occurs at a lower value than normal. As the punch-through short occurs, the metallisation will flow through the dielectric to create a low resistance short. However, in some instances where there is a particularly thin metallisation, such as 0.4 micron, or there is sufficient energy passed through the short, the metal will be vaporised and the short will clear but leave a cratered hole in the dielectric. Degraded performance may result but not a catastrophic failure. There is conjecture that the short in some circumstances might reappear or performance might continue to degrade.

Semiconductor junctions included in this classification are positive-negative (PN) junctions, P-type intrinsic N-type (PIN) junctions, and Schottky barrier junctions. Their sensitivity to ESD depends on geometry, size, resistivity, impurities, junction capacitance, thermal impedance, reverse leakage current and reverse breakdown voltage.

The energy required to damage a junction in the forward biased direction is generally ten times that required in the reverse biased direction. Emitter-base junctions in bipolar transistors, whether integrated circuit or a discrete transistor, are usually more susceptible to ESD damage than collector-base or collector-emitter junctions. This is primarily due to size and geometry, where the emitter-side wall experiences large energy-densities during reverse biased ESD. Because of larger areas, the collector-base and collector-emitter do not experience the same energy densities, although with the collector-base and collector-emitter it is possible to laterally forward bias the base-emitter. In this case, a current crowding at the emitter side will occur.

Junction field-effect transistors which have high impedance gates are particularly sensitive to ESD. They have extremely low gate-to-drain and gate-to-source leakage in the order of less than 1 nanoampere, and relatively high breakdown voltage of greater than 50 volts. Therefore, the gate-to-drain and gate-to-source are usually the most sensitive ESD paths. Figure 3 is a

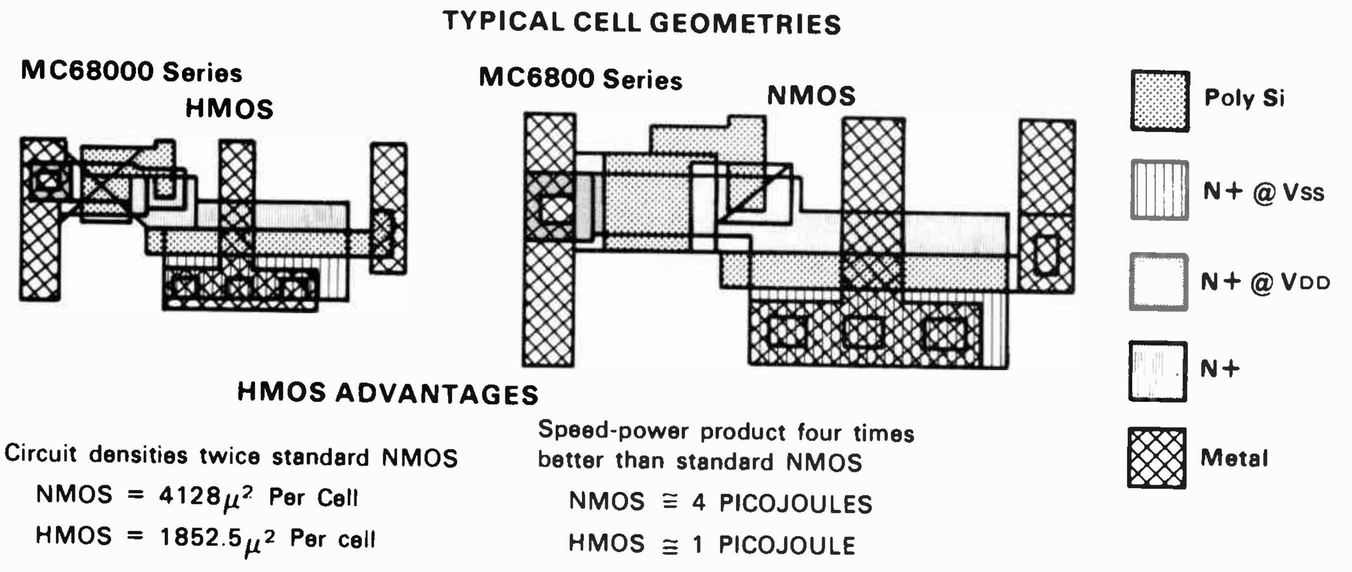


Figure 2. **Comparison of HMOS and NMOS technologies.**

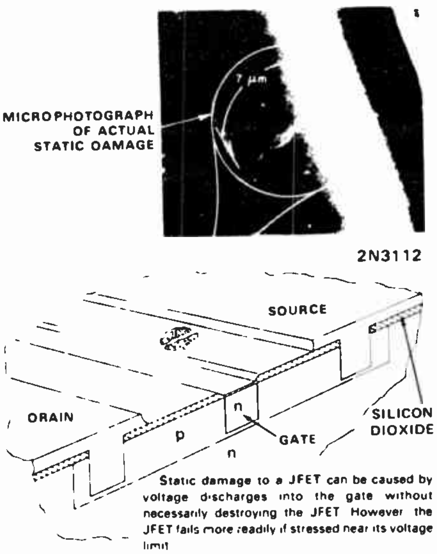


Figure 3. **Example of nondestructive ESD damage to a JFET device.**

classic example of non-destructive ESD damage to a JFET device. The device shown continued to function normally in the circuit. However, it experienced a dramatic decrease in its reverse breakdown voltage.

Schottky barrier junctions, such as the 1N57111 diode and TTL Schottky integrated circuits, are particularly sensitive to ESD because they have very thin junctions and the presence of metal increases the probability of ESD being carried through the junction.

Semiconductor junctions as sensitive ESD constituents are found not only in diodes, transistors, and bipolar integrated circuits, but also in MOS as parasitic diodes and input protection clamps. Although the input port junctions are meant to provide protection from ESD damage, the size of the protective junctions is limited due to cost and performance tradeoffs. Thus, ESD pulses of sufficient energy can damage the input protection junctions.

The temperature coefficient of extrinsic semiconductors is positive. That is, the higher the temperature, the higher the resistance. This feature prevents current crowding and hot spots from forming at low temperatures. However, in the reverse biased mode all the energy is being dissipated by the relatively large voltage drop across the relatively narrow depletion width of the junction. Due to geometrical effects, local resistance variations, and crystal defects, perfectly uniform current distribution does not occur across the junction. As an ESD occurs across the junction, the temperature at the depletion region increases quickly, and the extrinsic semiconducting material becomes an intrinsic semiconducting material, causing a sharp decrease in resistance which results in thermal secondary breakdown. The more rapid the discharge, the more uniform is the increase in temperature and therefore the current across the junction. This means that for short duration discharges of less than 10 nanoseconds, the resultant filament short is wide compared to longer duration discharges.

It is possible for spots to develop but not grow completely across the junction such that at low bias voltages they do not cause a failure condition. However, during operation at certain bias conditions, locally high current densities may exist with a correspondingly highly localised large increase in temperature at the previously formed hot spot locations, such that continued growth of a filament short may occur or silicon and metallisation may diffuse through the junction via the electromigration process at temperatures greater than 200°C. Low-leakage, high-breakdown JFET and Schottky barrier junctions seem to be particularly susceptible to this failure process.

It is this same failure process that requires the breakdown test of JFETs be performed as a leakage test rather than puting the junction into breakdown. With low-leakage junctions, highly localised currents can occur during junction reverse breakdown.

With the Schottky barrier junction, metallisation is immediately available to migrate through the junction at localised hot spots. As the current filament develops across a semiconductor junction, it is analogous to putting a parallel resistor across the junction of the same value as the short. However, in some marginally formed hot spots, it may be similar to putting a zener diode and a resistor in parallel with the junction. Failure indication of filament short from a high resistance short is high leakage.

Film Resistors: Resistor material adhering to an insulating substrate comes under the ESDS constituent classification of film resistor. The degree of sensitivity will depend on the ingredients and formulation of the resistor material and size-power considerations.

Hybrid microcircuits frequently contain either thin film resistors or thick film resistors. Hybrid designs which cannot tolerate large changes in resistance, such as precision voltage regulators, are sensitive to ESD.

Thick film resistors consist of a conductive metal oxide as the resistive element, a metal additive to improve electrical performance, and a glass frit to provide a support matrix, adhesion to the substrate, and resistivity control. Such parts are particularly sensitive to ESD. Since the charge is almost always negative for thick films, electrical discharge has been considered as a possible trimming method when conventional trimming overshoots the desired resistance tolerance. It has also been found that the thick film resistance changes are heavily dependent on voltage rather than energy.

Thin film resistors, on the other hand, are more energy dependent and do not have changes greater than 5% in resistance until the energy of discharge is sufficient to cause film rupture.

In addition to hybrid microcircuits, some monolithic integrated circuits may also contain encapsulated thin film resistors, such as polysilicon resistors, as part of an input protection circuit. Discrete encapsulated resistors which contain the film resistor structure are also sensitive to ESD.

Carbon film, metal oxide, and metal film resistors are somewhat sensitive to ESD, especially at low tolerance and low wattage ratings. A frequently recurring ESD problem with resistors is with the 0.05 W metal film, part RNC50, specified at 0.1% tolerance. Putting these parts in a polyethylene bag and rubbing them on another bag is sufficient to shift the tolerance of these resistors.

ESD failure mechanisms of film resistors are not well defined. This is partly the result of not knowing the ingredients and formulations of the resistor material, which are often held proprietary by the manufacturer.

For thick film resistors, the failure mechanism has been modelled as the creation of new shunt paths in a matrix of series-parallel resistors and infinitesimal capacitors isolating metallic islands. With the application of high electric fields, the dielectric breakdown of the glass frit or other isolating dielectric material is exceeded and the ensuing rupture welds metallic particles together in a conducting path known as metallisation melt. Since this model involves a dielectric breakdown process, it is mostly voltage dependent. ▶

WATCH THAT SOLDER SUCKER

Removing integrated circuits soldered directly to a printed circuit board usually requires sucking away the solder from a reheated pad or plated through hole. That is fine for bipolar circuitry, but it can be extremely dangerous for MOS devices. Dan Anderson, president of Anderson Effects, points out that standard plastic solder suckers have been found to produce a static surge of 5000 to 10 000 V at the tip. This tip is invariably in direct contact with a device's lead when the surge occurs, resulting in a damaged or destroyed device. Anderson Effects and other firms now offer static-free metalised plastic models that produce no static charge. For more information, contact Anderson Effects Inc., P.O. Box 657, Mentone, California 92359 USA.

It appears that the ESD behaviour of resistive materials is very much a function of the number of parallel current paths or the number of capacitive couplings between parallel paths in the film structure. The nature of the glass used in the material also appears to be quite important, both because it influences the distribution of the resistive elements and because it can act as a resistive element itself. Thus, the behaviour of different thick film resistor paths to ESD can vary greatly. ESD sensitivity testing, therefore, should be specified for critical tolerance thick film resistors.

For thin film resistors and encapsulated metal film, metal oxide, and carbon film resistors, the failure mechanism is primarily a thermal, energy-dependent process modelled as the destruction of minute shunt paths. This mechanism is associated with an increasing resistance shift on the thin film and metal film type resistor which appears to be voltage dependent. This negative shift is usually not more than 5% and is typically less than 1% before changing to positive shifts as ESD voltage increases.

Some thin film resistors, such as deposited tantalum nitride on SiO_2 substrates, may be so small and power-limited that ESD voltages greater than 5000 volts from a person can melt the resistor open. For most cases, however, a shift in resistance will be the failure indicator.

Thus, for circuit designs tolerant of large resistance changes, the failure may not be critical. Generally, after exposure to an ESD, the stability of the resistor is reduced and the degree of instability is directly related to the level of ESD. Temperature coefficient changes have been known to result from such ESD exposure.

For thick film resistors, the resistance shift is negative. The resistance change can easily exceed 50% with some thick film pastes. Some exceptions to this may occur, especially at low resistance values. For thin film, metal film, metal oxide and carbon film at lower ESD levels, small negative resistance shifts of less than 5% can be experienced. At higher ESD levels, large positive shifts greater than 10% can be experienced, depending on the power rating.

Metallisation Strips. Relatively narrow, thin metallisation strips on a substrate such as SiO_2, which carry current between terminals without any other energy-absorbing element in the path, are susceptible to ESD. These metallisations may consist primarily of aluminium or gold, but can also be multi-layered. The failure mechanism is burnout from joule heating. This type of constituent is often used in monolithic integrated circuits, hybrid microcircuits and multiple finger overlay transistor construction found in switching and high frequency transistors.

Joule heating is most likely to occur when: (1) the ESD source has very low contact resistance, resulting in high currents over short time constants, and (2) a low resistance large area diode is connected by the metallisation path between the two terminals, resulting in large currents due to the low voltage drop in the diode forward biased direction.

Increasing the width or thickness of the strip will decrease ESD sensitivity. The use of glassivation and thinner SiO_2 between the strip and the silicon also reduces ESDS. The failure indicator from this failure mode is open.

Passivated field-effect structures with nonconductive lids. Various NMOS and PMOS integrated circuit designs have been found to fail from very localised high concentrations of positively charged ions on the outer passivated surface of the die.

NMOS designs fail from excessive leakage currents as a result of field inversion between N+ junctions, such as thick field parasitic transistors, intermediate field parasitic transistors, EPROM transistors, and normal select transistors.

PMOS designs, such as the floating gate, EPROM or depletion type field effect transistors, fail when the negative charge on the floating gate is overcompensated by a positive charge, giving an erroneous unprogrammed indication. The effective field from the positively charged ions needed to create this inversion has been found to exceed 85 volts.

Hermetic packages which have recorded this failure mode have nonconductive lids made from non-transparent ceramic, transparent sapphire and transparent borosilicate glass.

These failures can be prevented by grounding the bottom surface of the lid over the die or by initiating preventive measures to avoid electrostatic charging of the nonconductive lid. This failure mechanism is most common with NMOS and PMOS UV-EPROMS having transparent lids. NMOS static random access memory (RAMs) in a ceramic package, however, have also been reported to fail from the ESD failure mechanism. Unless testing shows otherwise, any LSI integrated circuit with nonconductive lids could conceivably have field effect structures which are susceptible to failure from

CONDUCTIVE WRIST STRAP PROTECTS MICROCIRCUITS.

Royston Electronics has recently introduced a conductive wrist strap that meets the latest military specifications for quick release and resistance to line voltages from accidental contact.

Military users have specified these new requirements for wrist straps to prevent static electricity damage to microcircuits while radar, avionics, computer and other equipment is undergoing repair, maintenance or inspection in base stations or the field.

The CP401A grounding strap has a wrist attachment of "Velcro" hook-and-loop tape that separates with a slight pull for safety and to prevent a worker from breaking the grounding wire by inadvertently leaving the work area while still wearing the strap.

The wrist strap is made of conductive polyester ribbon for permanent conductivity, but with built-in resistance to protect the wearer against possible line voltages.

An alligator clip at the other end of the four-foot long grounding wire can be attached to any convenient ground, draining static electricity before it can build up to levels that are harmful to microcircuits.

The wrist strap and ground wire are joined by a standard snap fastener. An extra snap fastener provides a convenient connection for grounding electric tools, bench covers or other items in the work area that must be grounded to prevent static electricity build-up.

Information on this, and the comprehensive range of other anti-static devices, is available from Royston Electronics, Melbourne (03)543-5122 or Sydney (02)709-5293.

undesirable field inversion or gate threshold voltage shifting.

Failure mechanism involves positively charged ion clusters deposited on the die as a result of air breakdown in the air gap between the die surface and the bottom of the package lid. Charging of the bottom of the lid can be induced by several means, one of which is by freeze spraying the package with canned coolant. The positive charging rate of the freeze spray impinging on the top of the lid depends on the flow rate of the coolant from the can. At low flow rates, the charging is negative and does not induce failure; at high flow rates, sufficient positive charging can occur and induce failure. The localised air breakdown in the air gap of the package causes ionised streamers to form from the die to the lid. The positive charge on the bottom of the lid drives the positive charge in the streamer toward the die surface and attracts the negative charge toward the lid. This results in very localised clusters of positive ions on the die surface. Because of the nature of the air breakdown for certain package ambients, this charge is probably identical in type to the very large ions that can be experimentally created by positive corona discharge in the air.

These localised positive charges also cause the formation of inversion layer leakage paths between N+ diffusions and shift the gate threshold voltage on PMOS depletion type transistors. The formation of leakage paths and the gate threshold shifts give rise to isolated circuit failures. This failure mechanism is recoverable by neutralising the positive charge on the outer surface of the die.

On UV-EPROMs with transparent lids, recovery is nondestructive when 2737Å (2.737×10^{-7} metres) ultraviolet light with a minimum photon energy of 4.3 eV is applied to the chip for as short as three to five seconds.

Failure indicators for this failure mode come under the general classification of operational degradation. Operational degradation will take the form of a functional failure. In the case of NMOS UV-EPROMs, certain programmed bits appear unprogrammed and certain unprogrammed bits appear programmed. In one group of failure indicators, bit failures have been organised in columns where programmed bits appear unprogrammed. In another group of failure indicators, bit failures were organised on rows where unprogrammed bits appeared programmed.

The failure indicators for PMOS UV-EPROMs are random single-bit failures throughout the memory which would read as programmed but appear as unprogrammed. Failure indicators for NMOS static RAM have been reported as random hits stuck in "1" or "0" logic state and the adjacent cell also stuck but in the opposite logic state.

Piezoelectric crystal devices, such as quartz crystal oscillators and SAW devices, can fail from ESD, resulting in operational degradation. Electrical parameters of piezoelectric crystals contained within these parts are damaged by excessive driving current. Also, the piezoelectric effect from high voltages causes mechanical stress and movement to be generated in the crystal plate. When the voltage is too great, mechanical forces cause motion in excess of the elastic limits of the crystal and crystal fracture occurs. Fracture may occur as a lifted platelet, as has been experienced in lithium niobate SAW delay lines. Such fractures, when occurring in sufficient number, will cause enough change in the operating electrical characteristics to cause failure.

Closely spaced electrodes. When employing thick metallisation, such as 1.35 microns, gaseous arc discharge in an arc gap 50 microns wide can be used as a protection device to dissipate incoming high voltage spikes. In devices with closely spaced, unpassivated, thin electrodes, however, gaseous arc discharge can cause degraded performance.

Devices that employ thin, closely spaced electrodes include SAW devices. Other parts, such as high-frequency, multiple-finger transistors, and new technology, such as very large scale integration (VLSI) and very high speed integration (VHSI), could also be degraded to failure from arc discharge between metallisation runs. Arc discharge causes vaporisation and metal movement generally away from the space between the electrodes. Melting and fusing do not move the thin metal into the interelectrode regions, but the metal pulls together and flows or opens along the electrode lines. There can be fine metal globules in the gap region, but not in sufficient numbers to cause bridging. Shorting is not considered a major problem with unpassivated thin metal electrodes.

ESD failures have been experienced on SAW band pass filters with thin metal of 0.4 micron and electrode spacing of 3.0 microns.

Detecting ESD damage

By this time, hopefully, the reader has developed an acute sensitivity to the nature of ESD and the insidious nature of ESD damage to the function of an electronic device or a black box. Although there are many thousands of users worldwide, very few have the capability — that is, trained people and facilities — to perform the failure analysis that would lead to the recognition of ESD as the culprit in numerous equipment failures. ▶

NEW MATERIAL WILL OVERCOME SPACECRAFT LOSSES

A new composite material just successfully tested in Britain will overcome a problem in space that has led to the loss of at least two satellites.

Orbiting spacecraft are bombarded by high-energy electrons that cause electrostatic charges of up to 20 000 volts to build up on the surface of the craft. As a result a spontaneous electrical discharge can occur through the outer thermal protective material.

As well as damaging the covering of the satellite, the discharge can cause false electronic signals to disrupt the operation of the craft. If this should happen while it is being manoeuvred in space, the satellite could be lost.

British Aerospace says that these dangers will be eliminated by encasing the satellite in a new composite sandwich material which dissipates the surface charge the moment it hits the spacecraft. The material, which has been patented, also eliminates electrostatic-induced interference, minimises contamination and will prolong the operational life of satellites by maintaining the thermo-optical protection.

In tests, samples of the multi-layer thermal-insulating material have successfully withstood electron energy levels of 30 000 volts at intensities up to 30 times greater than those anticipated in space.

The material has shown that the electrostatic surface potential cannot build up to operationally dangerous levels even at temperatures as low as minus 170° Celsius, where the probability of a discharge is much greater.

British Aerospace says the new material is made up of one of two types of material already used for thermal insulation. However, in this new material it is arranged in composite sandwich form along with two conductive layers of aluminium or carbon, which are earthed to the satellite's main structure.

Electrons penetrating the outer skin are captured by the first conducting layer, while the second aluminium layer on the inner side of the material captures the more highly charged particles that may have penetrated further into the surface of the spacecraft. The inner conductive layer can also act as a radio frequency shield.

The multilayer technique may find further use in protecting spacecraft equipment such as thermal control mirrors, solar arrays and the back of antenna dishes, and British Aerospace say there may be other applications on the ground.

Even with trained people and the proper tools, identifying ESD damage can be difficult. Phil Kohlhaas of 3M Static Control Systems reported at a recent seminar (hosted by Warren Yates of Electronic Products magazine) that 3M sent 100 deliberately static-damaged devices to a testing laboratory. The lab performed a 100% failure analysis — SEM (scanning electron microscope), glass removal, metal removal, the works — and in 60% of the cases, could not identify ESD-related damage that had occurred.

ESD-induced failures are often mistaken for other types of failures. This is particularly true, according to Roy Walker of IITRI/RAC, when it comes to steady-state electrical over-stress failures. Agreeing with Walker, Hewlett-Packard's Kim Gray said he encountered a latch-problem in a CMOS device that appeared to result from steady-state-overstress failure; it turned out to be an ESD failure.

A lack of ESD awareness causes many people to limit ESD protection to only the most widely used susceptible devices — FETs without protection, and CMOS with double diode protection. But don't be lulled into a false sense of security if you're using bipolar devices. It's just more difficult to discern the ESD mechanism in a bipolar device than in a MOS device. Walker and others contend that there are many more ESD related problems in bipolar devices than we actually know about because of limitations of ESD failure analysis.

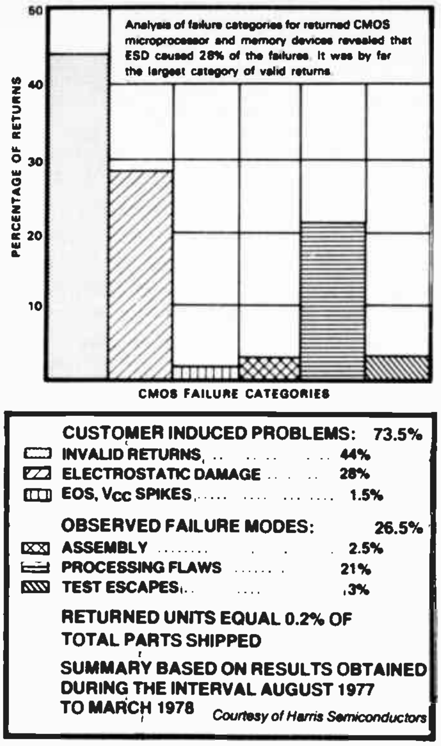


Figure 4. **Typical device degradation threshold.**

The ESD problem is big — make no mistake about that — even though it's really not possible to put a precise handle on just how big. For example, Gene Freeman presented some failure analysis data compiled by Harris Semiconductor on devices returned to them (see Figure 4). Note that ESD comprises the largest single failure mode.

Steve Halperin of Analytical Chemical Labs reported on his company's observations of equipment manufacturers. Where large boards of critical design are involved, he has seen up to a third of all boards started during a day enter a "repair and refurbish" function at some period during handling in the manufacturing facility. Cost of manufacturing failures can be prohibitive, but at least these types of problems are caught at the factory. But what about devices that are degraded by ESD but don't fail until later, out in the field? Halperin quoted figures from some computer manufacturers indicating that 70% of their field service calls were static related.

Degraded devices can become much more than just an expensive field service problem. We cannot ignore the possible substantial costs of product liability, as a failure in a critical end item system might mean substantial property losses or loss of life.

An effective plan to combat ESD requires a strong static awareness on the part of all concerned — factory assembly and test personnel, engineering, maintenance, and field service. But most of all, it requires a strong commitment on the part of top management.

Protecting your equipment

Once it is recognised that static discharge can degrade equipment per-

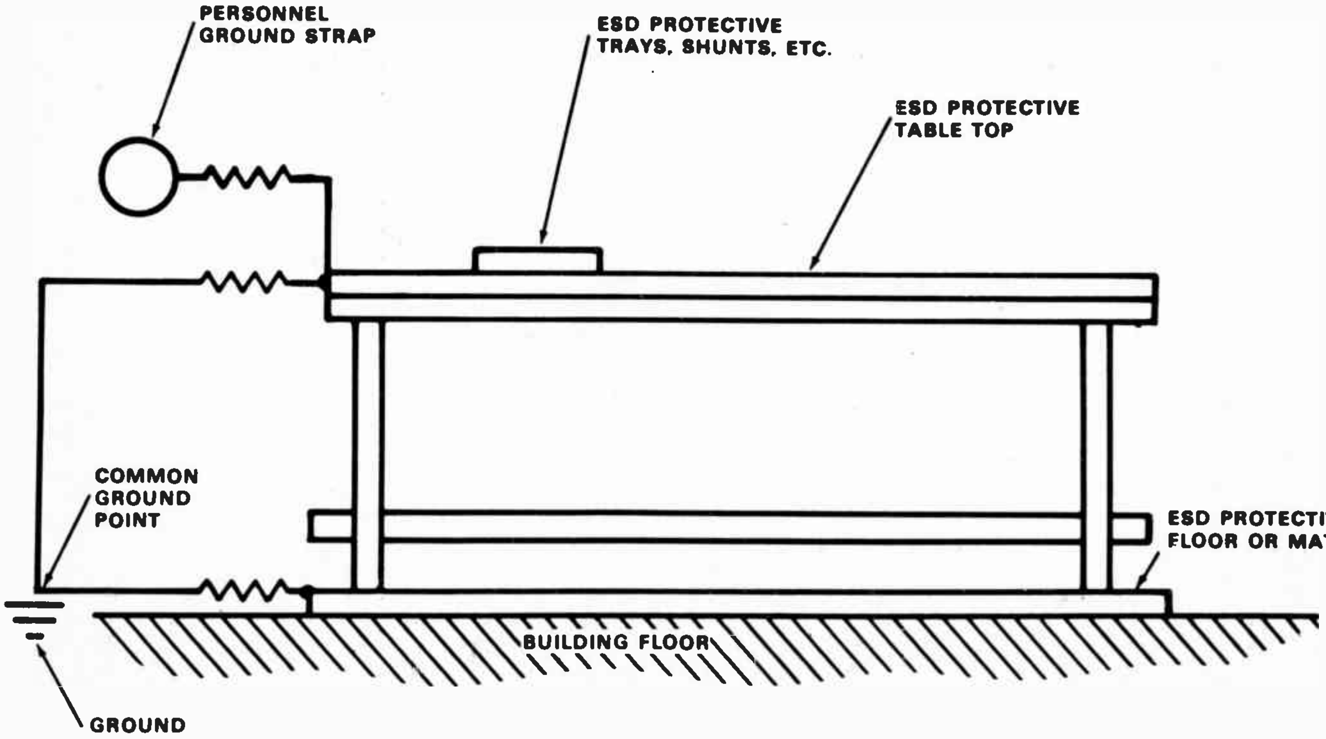


Figure 5. **Typical ESD grounded work bench.**

formance, and that in reality only the 'tip of the iceberg' can be identified as ESD, it is evident that ESD can be combatted only through protective measures. The first and most obvious key, as just stated, is top management's absolute commitment toward a total ESD program.

If management is astute enough to make this commitment, a total awareness and educational programme reaching to all individuals interfacing with the equipment is essential. After awareness is implementation, preparation of specifications and requirements to control work environments, identification and labelling of ESDS hardware, acquisition of antistatic handling equipment and work stations, and coordination of ESD programmes with both suppliers and users. ●

YOUR CHECK-LIST

1. **Identify static-sensitive parts. Manufacturers should be required to clearly mark all parts that are suspected to be sensitive to static charge. Markings should read "static-sensitive devices".**
2. **Provide procedural guidelines to all personnel involved in handling, packaging, testing, assembling, and reworking "static-sensitive devices".**
3. **Maintain good grounding techniques by keeping equipment and personnel at the same potential. Use conductive countertops, floor mattings, wrist straps, or arm sleeves, and make proper connections to a grounding source. (See Figure 5.)**
4. **Use conductive carriers for transporting, storing, and shipping static-sensitive parts.**
5. **Use neutralisers to neutralise charge on personnel, handling tools, and work surfaces.**
6. **Use a noncontacting static voltmeter to regularly monitor static charge in assembly area and on working personnel. This offers control and also keeps personnel static-conscious.**
7. **Failed parts should be treated with the same precautions; otherwise, the cause of the original failure may never be determined.**
8. **Keep all LRU assemblies stored away from high energy sources at all times (e.g: radar, laser, X-ray).**
9. **Keep connector caps on LRUs at all times whenever they are not installed. (Conductive caps are preferred.)**
10. **Never open an LRU on or remove an SRU unless at a properly equipped work station.**

Logic troubleshooting tools and techniques

Digital circuitry has had an explosive growth over the past decade and now pervades virtually every facet of electronics. Whilst the reliability of devices and equipment has improved dramatically during that period, things still do go wrong and equipment still needs servicing when it breaks down. 'Board level' servicing can solve problems quickly in the field, but economics demands those removed boards be repaired and recycled. This article details the problems encountered and the tools and techniques employed to fix them.

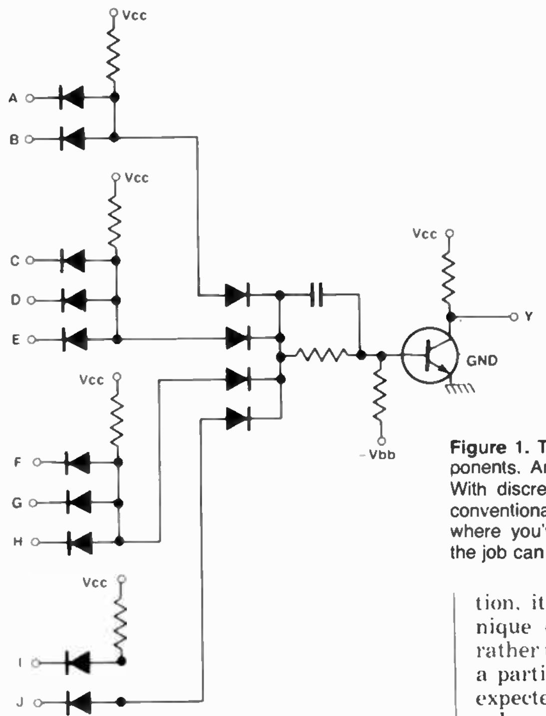
DIGITAL integrated circuits range from simple buffers and two-input gate packages through to complex purpose-built controllers and microprocessors. Finding faults in digital equipment requires a *fundamentally different* approach from fault-finding in analogue circuitry, where the multimeter and oscillator are the prime tools and component characteristics can be individually measured. In digital electronics, most 'components' are contained within the ICs — which are often multi-functional. Thus there is need for a different troubleshooting approach to be developed, and different tools used, based on the type of faults that develop and the 'signatures' they leave. Of necessity, this article does not cover microprocessor-based equipment — *that's* a whole subject on its own!

Faults & effects

When fault-finding circuits built from discrete components, the task is one of verifying relatively simple characteristics such as resistance, capacitance, or turn-on voltages of components with two or at most three nodes. (A 'node' is an active junction in a circuit, usually an input or an output.) While the function of the total circuit may be quite complex, each component in that circuit performs a relatively simple task and proper operation is easily verified.

In Figure 1, each diode, resistor, capacitor and transistor can be treated using a signal generator and a voltmeter, ohmmeter, diode checker or oscilloscope — the traditional servicing tools. But when this circuit is built in integrated circuit form, these components are no longer accessible. It now becomes necessary to test the operation of the complete circuit function.

Thus an important difference between discrete circuitry and circuits built from digital ICs is in the complexity of the functions performed by these 'components'. Unlike the resistor, capacitor, diode or transistor, which must be interconnected to form a circuit function, digital ICs perform complete, complex functions. Instead of observing simple characteristics, it is now necessary to observe complex digital signals and decide if these signals are correct according to the function the IC is meant to perform.



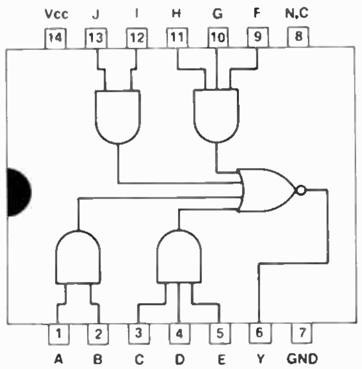


Figure 1. This 10-input gate is made up from 23 components. An IC to do the same job is shown at right. With discrete components, finding faults is easy with conventional meters, etc. When it's all inside an IC, where you've only got access to inputs and outputs, the job can be much harder. But there are ways.

Verifying proper component operation now requires 'stimulating' and observing many inputs (in Figure 1 there are 10 inputs) while simultaneously observing several outputs (up to two or three and at times as many as eight).

Thus another fundamental difference between circuitry built from discrete components and digital ICs is the number of inputs and outputs associated with each component, and the need to stimulate and observe these simultaneously.

In addition to the problems of simultaneity of signals and complexity of functions at the component level, the digital IC has introduced a new degree of complexity at the circuit level. Circuits which perplex all but their designer are commonplace. Given enough time, these circuits can be studied and their operation understood, but this is not an affordable luxury for those involved in troubleshooting electronic circuits. Without understanding a circuit's intricate operation, it becomes necessary to have a technique of quickly testing each component rather than attempting to isolate a failure to a particular circuit segment by testing for expected signals.

In order to solve these problems and make fault-finding of digital circuits more efficient, it is necessary to take advantage of the digital nature of the signals involved. Tools and techniques designed to service analogue circuits do not take advantage of this digital nature and thus are less efficient when used to troubleshoot digital circuits.

Figure 2 shows a typical TTL (Transistor-Transistor-Logic) signal. This might as well be any analogue signal when viewed on an oscilloscope. The oscilloscope displays absolute voltage with respect to time, but in the digital world absolute values are unimportant.

A digital signal exists in one of two or three states — high, low and undefined or in-between level — each determined by a threshold voltage. It is the relative value of the signal voltage with respect to these thresholds that determines the state of the digital signal, and this digital state determines the operation of the IC, not absolute levels.

In Figure 2, if the signal is greater than 2.4 volts, it is a high state and it is unimportant whether the level is 2.8 or 3.0 volts. Similarly for a low state the voltage must be below 0.4 volts. It is not important what the absolute level *is* as long as it is below this threshold. Thus when using an oscilloscope, the serviceman must over and over again determine if the signal meets the threshold requirement for the desired digital state.

Within a digital logic family, such as TTL, the timing characteristics of each component are well defined. Each gate in the TTL logic family displays a characteristic propagation delay time, rise time and fall time. The effects of these timing parameters on circuit operation are taken into account by the designer. Once a design has been developed beyond breadboard or prototype stage and is into production, problems due to design have (hopefully) been corrected.

An important characteristic of digital ICs is that when they fail, *they fail catastrophically*. This means that timing parameters rarely degrade or become marginal. Thus, observing on an oscilloscope and making repeated decisions on the validity of timing parameters is time consuming and contributes very little to the fault-finding process. Once problems due to design are corrected, the fact that pulse activity *exists* is usually enough indication of proper IC operation without further observation of pulse width, repetition rate, rise time or fall time.

Figure 3 shows a problem created by the TTL logic family. The output stage of a TTL device is a transistor totem pole. In either the high or low state, it is a low impedance. In the low state it is a saturated transistor to ground. It thus appears as 5-10 ohms to ground. This presents a problem to in-circuit stimulation.

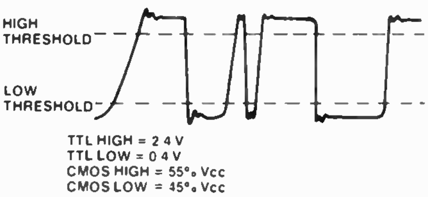


Figure 2. Digital circuits work on 'thresholds', and signals must be 'above' or 'below' the given high and low thresholds, which are different for the different 'families' — CMOS and TTL.

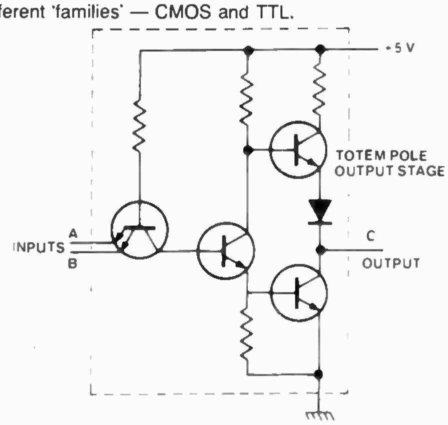


Figure 3. TTL ICs have 'totem pole' output stages, as shown above. When attempting to 'stimulate' an output node, such as C above, it is necessary to override the low impedance output stage, which consists of a saturated transistor.

A signal source used to inject a pulse at a node which is driven by a TTL output must have sufficient power to override the low impedance output state. If you use a squarewave signal generator for fault-finding it must provide this capability, otherwise it is necessary either to cut printed circuit traces or pull out IC leads in order to stimulate the circuit being tested. Both of these practices are time consuming and lead to unreliable repairs.

Thus the use of the traditional oscilloscope and the traditional signal sources is inefficient. Since the diodes and transistors are packaged in the IC, use of diode checkers is also marginal, if not impossible.

These tools are general purpose tools that can be applied to any situation if you have enough time. But with the quantity and complexity of today's electronic circuits, it makes sense to find the most efficient solution to the problem at hand. This suggests using the oscilloscope, diode checkers and voltmeters on analogue circuits where they really shine, and using instruments that take advantage of the digital nature of signals on the digital circuitry to be repaired. We'll get to them a little later.

In order to repair digital equipment efficiently, it is important to understand the type of failures found in digital circuits. These can be categorised into two main classes — those caused by a failure *internal* to an IC and those caused by a failure in the circuit *external* to the IC.

Four types of failures can occur internally to an IC. These are (1) an open bond on either an input or output, (2) a short between an input or output and Vcc or ground, (3) a short between two pins (neither of which are Vcc or ground), and (4) a failure in the internal circuitry (often called the steering circuitry) of the IC.

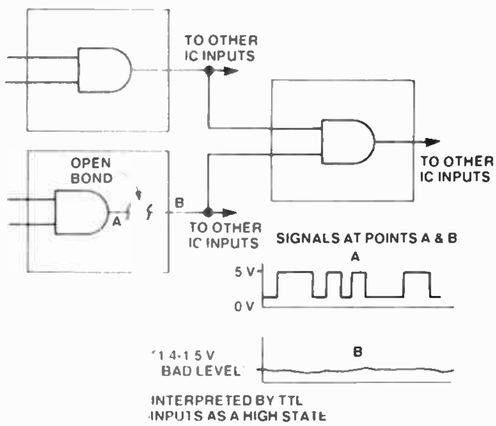
In addition to these four failures internal to an IC, there are four failures that can occur in the circuit external to the IC. These are (1) a short between a node and Vcc or ground, (2) a short between two nodes (neither of which are Vcc or ground), (3) an open signal path, and (4) a failure of an analogue component.

Before showing how to detect each of these failures we will discuss the effect each has upon circuit operation.

The first failure (internal to an IC) mentioned was an open bond on either an input or output. The failure has a different effect depending on whether it is an open output or an open input bond. In the case of an open output bond (Figure 4), the inputs driven by that output are left to float. In TTL circuits a floating input rises to approximately 1.4 to 1.5 volts and usually has the same effect on circuit operation as a high logic level. Thus an open output bond will cause all inputs driven by that output to float to a bad level since 1.5 volts is less than the high threshold level of 2.0 volts and greater than the low threshold level of 0.4 volt. In TTL a floating input is interpreted as a high level. Thus the effect will be that these inputs will respond to this bad level as though it were a static-high signal.

In the case of an open input bond (Figure 5), we find that the open circuit blocks the signal driving the input from entering the IC

Figure 4. An open output bond allows all inputs driven by that output to float to a 'bad level', usually interpreted as a high. Signals at points A and B illustrated below. ▼



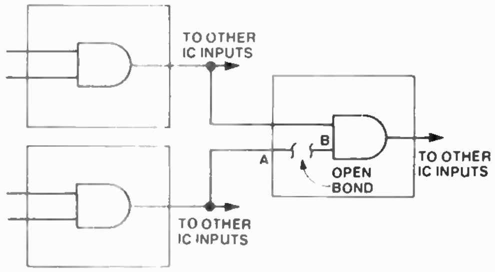


Figure 5. An open input bond blocks the incoming signal, allowing the input to float to a 'bad level' — interpreted as a high. Signals at points A and B illustrated above.

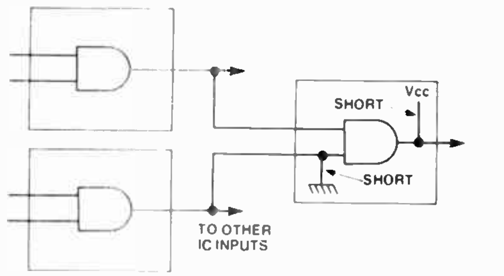


Figure 6. When you get an internal short to ground, the affected node is always pulled low. When shorted to Vcc (supply), the affected node is always pulled high.

chip. The input on the chip is thus allowed to float and will respond as though it were a static high signal. It is important to realise that since the open circuit occurs on the input inside the IC, the digital signal driving this input will be unaffected by the open circuit and will be detectable when looking at the input pin (such as at Point A in Figure 5). The effect will be to block this signal inside the IC and the resulting IC operation will be as though the input were a static high.

A short between an input or output and Vcc or ground has the effect of holding all signal lines connected to that input or output either high (in the case of a short to Vcc) or low (if shorted to ground) (Figure 6). In many cases, this will cause expected signal activity at points beyond the short to disappear, and thus this type of failure is catastrophic in terms of circuit operation. ▶

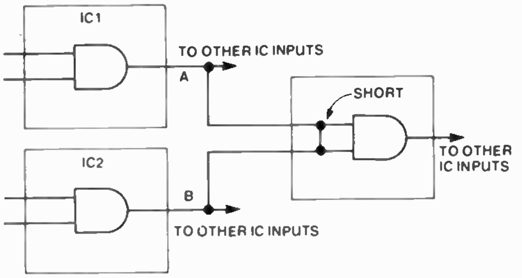
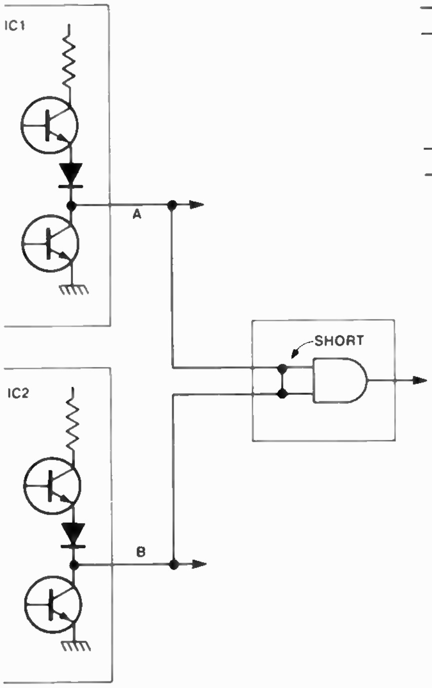


Figure 7. A short between two inputs makes a low-going driver pull the other driver low too. In IC2 at right, if B is low, it is pulled low by a saturated transistor, which pulls A low too. ▲ ▶



A short between two pins is not as straightforward to analyse as the short to Vcc or ground. When two pins are shorted, the outputs driving those pins oppose each other when one attempts to pull the pins high while the other attempts to pull them low (Figure 7). In this situation the output attempting to go high will supply current through the upper saturated transistor of its totem pole output stage, while the output attempting to go low will sink this current through the lower saturated transistor of its totem pole output stage. The net effect is that the short will be pulled to a low state by the saturated transistor to ground. Whenever both outputs attempt to go high simultaneously, or to go low simultaneously, the shorted pins will respond promptly. But whenever one output attempts to go low the short will be constrained to be low.

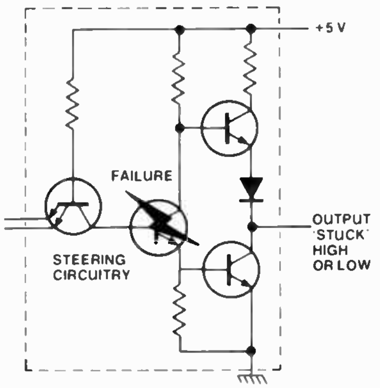


Figure 8. An internal failure of the steering circuitry will either cause the output to be 'stuck' high or low.

The fourth failure internal to an IC is a failure of the internal steering circuitry of the IC (Figure 8). This has the effect of permanently turning on either the upper transistor of the output totem pole, thus locking the output in the high state, or turning on the lower transistor of the totem pole, thus locking the output in the low state. Thus this failure blocks the signal flow and has a catastrophic effect on circuit operation.

A short between a node and Vcc or ground external to the IC is indistinguishable from a short internal to the IC. Both will cause the signal lines connected to the node to be either always high (for shorts to Vcc) or always low (for shorts to ground). When this type of failure is encountered a very close physical examination of the circuit may reveal if the failure is external to the IC, but it can be determined using 'pulsing' and 'tracing' tools, as explained later.

An open signal path in the circuit has a similar effect to an open output bond driving the node (Figure 9). All inputs to the right of the open will be allowed to float to a bad level and will thus appear as a static high level in circuit operation. Those inputs to the left of the open will be unaffected by the open and will thus respond as expected.

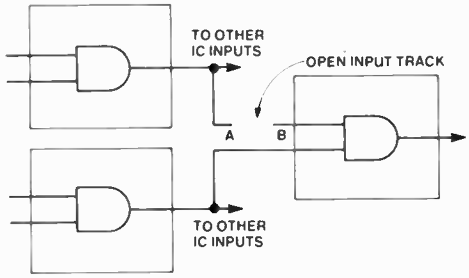


Figure 9. An open input track (external to the IC) has a similar effect as an open driver output bond. B will float to a bad level while A will still have signal on it.

The problem of open-collector outputs — 'wired-ANDs', 'wired-ORs' — is different from the other cases described. Open-collector outputs differ in that they do not have an active logic-high current source. Instead, the output stage collector (Q3 in Figure 10a) is left unconnected. Thus the output stage can *sink* current in a logic low state, but cannot *source* any current in the high state. This is provided by the 'pullup' resistor R_L. Generally, you will find several open-collector gates are interconnected in parallel, as shown in Figure 10b.

So long as every output stage is turned off, the voltage at the common connection node is near Vcc, but when *any one* gate output is driven on, the node voltage drops to the low state (near 0 V). The common node thus acts as an AND gate in itself (hence 'wired-AND'). This circuit is 'wired-NAND' in TTL circuits if the inputs and outputs are active low. In other families, it's an OR function. When looking for faults here, the output has to be looked at in conjunction with the input.

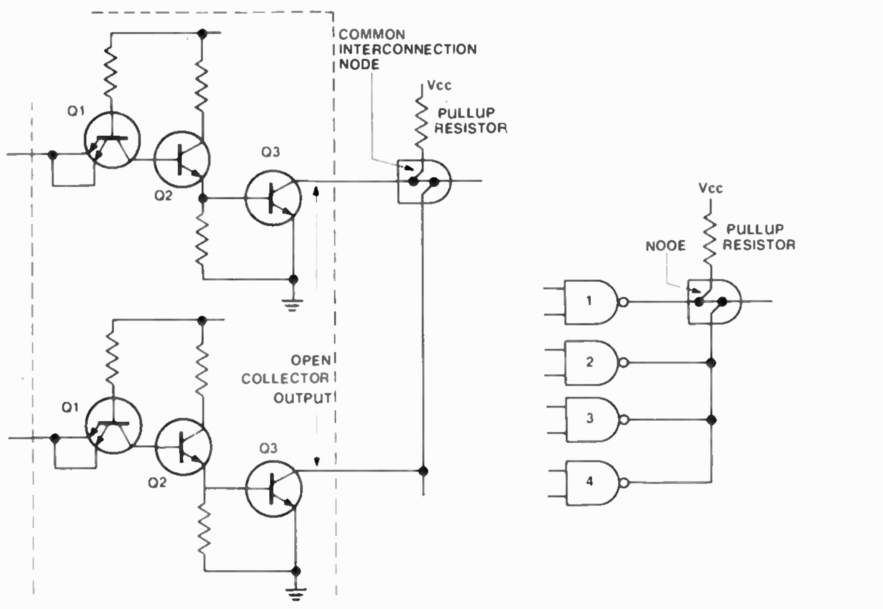


Figure 10. The 'open collector problem'. Open collector output stages do not have an active high source, this being provided externally by a 'pullup' resistor. When gates are connected in the 'wired-AND', 'wired-OR' configuration, the output of one IC can constratin the node to be in a state other than that defined by the gates' truth table.

ACKNOWLEDGEMENT

We would like to acknowledge the assistance kindly provided by the Instrument Group of Hewlett-Packard Australia Limited.

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Logic troubleshooting

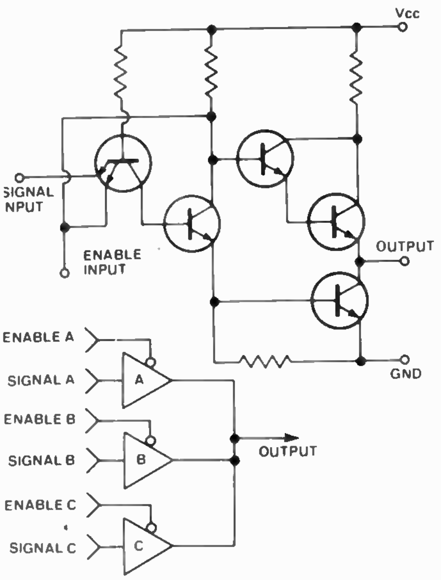


Figure 11. The internal circuitry of a 'tri-state' driver is shown above. The output can be high, low or open-circuit. Outputs are generally wired in parallel, as in the wired-AND circuit.

A now widely used logic type, known as 'tri-state' logic, is a development of this idea. It is found extensively in microprocessor equipment. It is particularly found in *bussed* systems where a multitude of devices might share a common, multi-track buss. Figure 11a shows the general internal circuitry of tri-state logic. The output can be high, low or (virtually) open circuit. The control input determines whether the output is 'enabled' (i.e: operative) or not. The outputs of tri-state logic are wired in parallel, sharing the same line. Only one driver is enabled at a time. It operates in a similar fashion to the wired-AND, the difference being that an 'enable' signal *must* be present for a particular gate's output to be active, otherwise the outputs remain in the open state.

Tools

There are two fundamental classes of tools employed in logic troubleshooting — *'stimulators'* and *'indicators'*. A stimulator is nothing more than a relatively simple pulse signal generator that is used to 'stimulate' sections of a faulty circuit into action. Hence the generic name. However, they are often called simply 'pulsers'. An indicator is just that — a device that will indicate the state of a point in a digital circuit: whether it's high, low, sitting at a bad level or pulsing. Generally a LED or other light indicator is used to signal the condition, some makes using several, coloured indicators.

There are two types of indicators — the *contact* type, where a connection is actually made to the circuit, and the *non-contact* type which picks up currents flowing in the circuit interconnections (generally called 'current tracers').

For user convenience, these tools are generally built into handheld, pen-sized cases with a sharp 'probe' point at the business end. Hence they are generally known as probes — *pulser probes* for the stimulators, *logic probes* for the indicators (because they indicate logic states). Actually, the use of the word 'probe' here is a misnomer. A probe is in fact a blunt instrument. Examine a logic probe, and except for the current tracer type you'll find they have sharp points!

Since it is necessary to observe dynamic signal activity, as well as the static levels, logic probes usually have pulse stretching circuitry that can detect pulses as narrow as 10 ns and stretch them so that a readily visible blink can be seen. Thus if a low signal pulses high, the logic probe will blink 'on'. If a high signal pulses low, the probe will blink 'off'.

With some logic probes, a pulse memory may be provided. This enables the probe to monitor a signal line for single shot or low frequency pulses over extended periods of time. If a pulse occurs, this will be indicated by the device, which will remain 'on' until reset by the user.

The existence of a pulse train is indicated by flashing the lamp indicator at a constant rate (typically 10 Hz) when a pulse train is present.

Thus a logic probe enables you to view static (high or low) logic levels, single-shot pulses and pulse trains. Automatic threshold detection is generally included as it eliminates the need to determine repeatedly whether a signal is above or below the threshold, and can be employed to show open-circuit conditions also. A TTL/CMOS switch is a necessity so that a probe can be used on both device families. Some makes work on both families, without a switch.

Current tracer logic probes require no contact with the circuit at all. At the business end of a current tracer probe is a small magnetic pickup, generally consisting of a coil wound on a tiny core which has a split, permitting any external field to induce currents into the winding. Of necessity, it works on pulse (ac) signals, detecting *current change*, not dc levels. Pulses are stretched so that the display, usually a LED, can be easily seen. Sensitivity can be arranged so that the current tracer will detect the current it takes to charge the gate input capacitance of CMOS devices.

Current tracers are very useful in sorting out 'stuck' nodes, particularly where there are many elements common to the node and too few ways to isolate the one bad component. It can be done, too, without cutting pc board tracks or lifting IC pins. They are also very useful in tracing signals on multi-layer boards.

Poke & peek

The mainstay of all digital troubleshooting is stimulus-response testing. It is necessary to apply a signal and observe the response to determine if the device is operating properly. As was pointed out earlier, this can sometimes be very difficult to do in TTL circuitry.

A logic pulse provides the solution. It is used to inject into the circuit a single pulse of proper amplitude and polarity — *forcing* something to happen. If the node happens to be low, it will be pulsed high, and if high, it will be pulsed low.

Generally, logic pulsers are capable of supplying both continuous pulse trains and single-shot pulses.

A logic pulser used in conjunction with a current tracer probe is particularly useful in tracing supply rail short circuits and stuck nodes having many common elements.

These tools are useful in troubleshooting both sequential logic circuits (counter, timer and simple control systems, etc) and parallel bit circuitry (microprocessor systems, etc). However, in parallel systems which are partially working it becomes necessary to see the simultaneous action of many lines or nodes, and a more complex technique, called *signature analysis*, is necessary. A signature analyser is the appropriate tool hre, and they come in many forms. Signature analysis, though, is a whole subject in itself and we'll have to leave that to another time.

Techniques

Your first 'port of call' should *always* be the power supply, particularly in the case of total collapse. If the power supply itself proves OK, but the rails on the pc board show a volt or less, then shorted rails should be suspected. If the supply rails are healthy, then the very next step is to attempt to narrow down the malfunctioning area as much as possible by examining the observable characteristics of the failure. Try to localise it to a circuit section or to as few sections as possible. Then you can proceed to eliminate circuit components step by step by looking for improper key signals between circuits — which is where the logic probe and current tracer come into their own. Table 1 details the general run of faults and how to detect them using the stimulus-response technique.

Dependence upon a well-written service manual is the key to this phase of troubleshooting. Isolating a failure to a single circuit requires knowledge of the instrument or system and its operating characteristics. A well-written manual will indicate key signals to be observed. The logic probe will provide a rapid means of observing the presence of these signals.

Once a failure has been isolated to a single circuit, the tools described can be used to observe the effect of the failure on circuit operation and to locate the failure to its cause (either an IC or a fault in the circuit external to the IC).

The logic probe is used to observe the signal activity on inputs and to view the resulting output signals. From this information, a decision can be made as to the proper operation of the IC.

For example, if a clock signal is occurring on a decade counter and the enabling inputs (usually reset lines) are in the enabled state, then the output should be counting. A logic probe will allow the clock and enabling inputs to be observed, and, if pulse activity is indicated on the outputs, then the IC can be assumed to be operating properly.

As stated before, usually it is not necessary to see the actual timing of the output signals, since ICs fail catastrophically. The occurrence of pulse activity is often enough indication of proper operation.

When more detailed study is desired, or when input signal activity is missing, the logic pulser can be used to inject input sig-

TABLE 1.

FAULT	INDICATOR	STIMULUS	TEST METHOD
Shorted node	Current tracer	Pulser or circuit signals(1)	• Pulse node • Follow current pulses to short
Stuck data buss	Current tracer	Pulser or circuit signals(1)	• Pulse buss line • Trace current to device holding the buss line in a stuck condition.
Signal line short to Vcc or ground	Logic probe and/ or current tracer	Pulser	• Pulse and probe test point simultaneously • Short to Vcc or ground cannot be overridden by pulsing • Pulse test point, and follow current pulses to the short with tracer
Vcc to ground short	Current tracer	Pulser	• Remove power from test circuit • Disconnect electrolytic bypass capacitors • Pulse across Vcc and ground • Trace current to fault
Suspected internally open IC	Logic probe	Pulser or circuit signals(1)	• Pulse device input • Probe output for response
Solder bridge	Current tracer	Pulser or circuit signals(1)	• Pulse suspect line(s) • Trace current pulses to the fault (Light goes out when solder bridge passed)

1. Use the pulser to provide stimulus, or use normal circuit signals, whichever is most convenient.

nals, and the probe used to monitor the response. This technique is especially good when testing gates and other combinatorial devices. A logic pulser can be used to cause the inputs to go to a state which will cause a change in the output state.

For example, a three-input NAND gate which has high, low, low inputs will have a high output. By pulsing the two low inputs high using a logic pulser, the output will pulse low, and can be detected by a logic probe. This then indicates that the IC is operating properly.

A logic pulser is also valuable for replacing the clock in a digital circuit, thus allowing the circuit to be single-stepped while the logic probe is used to observe the changes in the circuit's state.

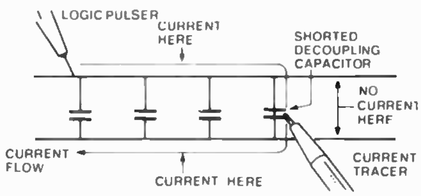
The first step might be called the 'mapping' step, since the effect is to *map out* the problem areas for further investigation. It is important to do a complete 'mapping' of the circuit before proceeding to analyse each of the indicated failures. Prematurely studying a fault can result in overlooking faults which cause multiple failures, such as shorts between two nodes. This often leads to the needless replacement of a good IC and much wasted time. With a complete trouble-area 'map' you can begin to determine the type and cause of the failures. This is done by systematically eliminating the possible failures, as discussed earlier.

The first failure to test for is an open bond in the IC driving the failed node (the Figure 4 problem). A logic probe provides a quick and accurate test for this failure. If the output bond is open, then the node will float to a bad level. By probing the node, the logic probe will quickly indicate a bad level. If a bad level is indicated, then the IC driving the node should be replaced and retested.

If the node is not a bad level, then a test for a short to Vcc or ground should be made next (the Figure 6 problem). This is best done using a logic pulser and current tracer. The problem is to determine if the driver is dead, or if a shorted input is clamping the node to a fixed value.

Use a logic probe and pulser to test the node's logic state and to see if the state can be changed (shorts to Vcc or ground cannot be

Figure 12. Tracing a supply rail short.



Use the probe and pulser to test the node's logic state and to see if the state can be changed (shorts to Vcc or ground cannot be overridden by pulsing). By pulsing the node you can follow the current directly to the faulty input using a current tracer (Figure 13).

If the node is shorted to Vcc or ground there are two possible causes. The first is a short in the circuit external to the ICs and the other is a short internal to one of the IC's attached to the node. The external short should be detected by an examination of the circuit. If no external short is found, then the cause is equally likely to be any one of the IC's attached to the node. The only suggestion that can be made (based on experience) is to first replace the IC driving the node, and if that does not solve the problem try each of the other IC's individually until the short is eliminated. (It might be noted that on occasion analogue components such as resistors or capacitors attached to the node have shorted.)

If the node is not shorted to Vcc or ground, nor is it an open output bond, then we should look for a short between two nodes. This can be done in one of two ways. First the logic pulser can be used to pulse the failing node being studied, and the logic probe can be used to observe each of the remaining failing nodes. If a short exists between the node being studied and one of the other failing nodes, then the pulser will cause the node being probed to change state (i.e: the probe will detect a pulse). To ensure that a short exists, the probe and pulser should be reversed and the test made again.

If the failure is a short there are two possible causes. The most likely is a problem in the circuit external to the ICs. This can be detected by physically examining the circuit, but shorts are not always obvious if only

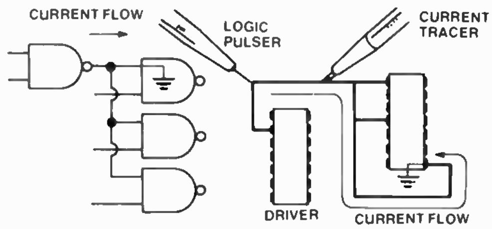


Figure 13. Tracing a stuck node. ▲

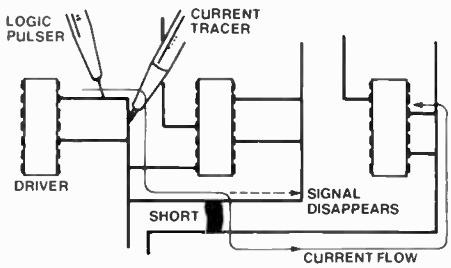


Figure 14. Tracing a track short.

traced down to an area. A current tracer is best to pinpoint a short between tracks by tracing current from the pulser. When the short is passed, the signal disappears (see Figure 14).

If the two nodes which are shorted are common to one IC, then the failure must be internal to that IC (the Figure 7 problem). If after examining the circuit no short can be found external to the IC, then the IC should be replaced.

If the failure is not a short between two nodes, then there are only two possibilities left. They are that the failure is an open input bond or a failure of the internal circuitry of the IC (Figures 5 and 8 problem). In either case, this IC should now be replaced. Thus by systematically eliminating the IC failures, the cause can be located.

An important step at any point where an IC is replaced is the retesting of the circuit. If the testing again indicates a failure, then more study of the problem must be made with the knowledge that the failure is not in the IC that has just been replaced.

An open track on the pc board (the Figure 9 problem) is best located with a logic probe, using either circuit signals or a pulser to provide the stimulus. The logic probe provides a rapid means of not only detecting but also physically locating the open.

Since an open signal path allows the input to the 'right' of the open to float to a bad level, the logic probe can be used to test the input of each IC for a bad level. Once an input floating at a bad level is detected, the logic probe can be used to follow the circuit back from the input looking for the open. This can be done because the circuit to the 'left' of the open will be a good logic level (either high, low or pulsing), while the circuit to the 'right' will be a bad level, precisely locating the open. The open can then be repaired and the circuit tested.

This systematic elimination of possible failures in digital circuits by the use of such special tools will ensure a rapid and accurate repair. Because these instruments provide a digital solution to the digital problem, improvements in servicing time of at last 4:1 are easily achieved over the use of analogue instruments. ●

Beating the RS232 blues

A serial interface should be the simplest way to connect two pieces of computer gear together. Unfortunately, RS232 complicates matters.

Grahame Wideman

ONLY TWO pieces of wire are needed to allow one computer device to talk to another, and three if you want a two-way conversation. So you would think that hooking together computer equipment with serial interfaces would be easy — provided, of course, that the various equipment manufacturers had adopted a standard for their interfaces. And herein lies a problem.

The 'standard' which was adopted for serial interface was one known as 'RS232'. RS232 is a standard of the American Electronics Industries Association, and was originally intended for the interface between 'Data Terminal Equipment' (DTE — in other words a computer 'dumb' terminal) and 'Data Communications Equipment' (DCE — equipment which facilitates communication to a remote computer, like a modem).

The standard specifies the electrical characteristics of the interface signals, along with the shape and pin assignments of the connectors to be used. In addition there are certain other conventions which go along with this standard, like the commonly used data rates and formats.

Now, although it's possible to borrow the electrical and timing conventions from this standard, many aspects are ambiguous. As mentioned above, RS232 specifies two different 'sexes' of equipment, terminal equipment and communications equipment, each with their own sex of connector, and their own connector pin assignments. But the standard is now being applied also to computers, printers, plotters, digitising tablets, speech synthesisers and so on, which don't fall conveniently into the category of either sex. Consequently any particular piece of equipment has a more or less arbitrary sex assignment.

Furthermore, RS232 contains specifications for using its connectors and signals for a large number of different applications. Since today's equipment needs only the simplest of such arrangements, most of RS232's features are not used, and in fact merely add to the confusion as manufacturers arbitrarily select the few features they need for their interface.

OK, the fact that the interface is somewhat arbitrary on any particular piece of equipment would be compensated if the equipment manual told you how it worked. Not the case. In fact the description of how the RS232 interface works is *almost universally the worst described part of the manual,* ranging from extremely ambiguous to downright wrong.

This month I am going to describe the theory of how RS232 is supposed to work. In a following issue I will describe a test unit which will patch any two devices together, and monitor what they are saying. You may wish to build one, or borrow the principles to understand how to test an interface by some alternative method.

The basics of a serial interface

There are many possible ways to make a serial communications 'channel'; RS232 is just one method. Let us examine serial interfaces in general, and see how RS232 implements the various features involved.

I should point out here that many of these features are not strictly a part of RS232, but are conventions which are used with it. The best way to declare something as a 'convention' is by referring to data on the ICs used to implement RS232 serial interfaces, namely the 'UART' which formats the data (such as the National 5303 and similar), and the 'line driver' and 'receiver' which actually send and receive the electrical signals on the serial cable (National LM1488 and 1489 respectively).

Suppose we are dealing with the simplest type of interface, one in which there is a 'sender' and a 'receiver', such as may be the case where a computer sends data to a line printer. Two wires connect the two devices, one wire being 'Ground' or zero volts, the other wire carrying the data.

Ones and zeros

The first task is to decide how to represent the binary 'one' and 'zero' as voltages. A TTL logic IC regards a voltage less than 0.4 V as a logic zero, and a voltage greater than 2.8 V as a logic one. A TTL output is not, for various reasons, suited to sending data down a long wire, so RS232 does things differently. A 'zero' is represented by a 'high' voltage between +3 V and +12 V (for some reason also called 'space'), while a 'one' is represented by a 'low' voltage between −3 V and −12 V (also called 'mark'). The range between −3 V and +3 V is undefined.

Next we must decide in what order and with what timing the bits are to be sent down the wire. RS232 calls the unit of data transmission a 'character', even though the data sent may not actually represent a character. A particular device may be set to transmit or receive 5, 6, 7 or 8-bit characters, with seven being the most common (because seven bits will represent the entire ASCII set of 128 characters), and eight the next most popular. These characters are sent *least significant bit first.* Using the scheme as I have so far described it, the letter 'B', which is ASCII 42 hexadecimal, or 66 decimal, would appear on the line as (see also Figure 1):

High low high high high high low
(7-bit code)

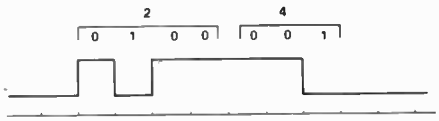


Figure 1. **The letter 'B' (hex 42) represented as a sequence of voltage levels, as used by RS232 devices.**

How does the receiver know when a particular character starts? We could use a third wire to signal that a character is starting on the second wire. This is a form of 'synchronous' communication, and is not used with any personal computer equipment. Instead RS232 has a way of telling the receiver that a character is starting. It works as follows.

Start bit

Suppose the receiver receives the above letter 'B'. Normally the communications line sits at 'mark' or low. Along comes bit one, which is a high, and immediately the receiver knows a character is coming in. Now, assuming that the receiver and sender are set so that they agree as to how long each bit is, the receiver will be able to recognise a high, then a low, then another low and so on, until the whole 'B' has been received.

However, suppose that instead the letter 'A' was sent, which is 41 hex, and therefore is represented (also see Figure 2) as:

Low high high high high high low

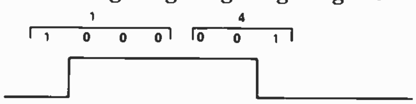


Figure 2. **The letter 'A' (hex 41) represented in RS232 voltage levels.**

This time, by the time the receiver finds out something is happening, it's already on the second bit! And what if you had a character composed entirely of lows?

The way around this problem is to prefix every character with a 'start' bit, which is invariably high.

Stop bit

This still leaves one problem. Suppose we send several hundred characters in a row. It would be unreasonable to expect that the sender and receiver agree as to the time-per-bit to such great accuracy that they would still be in step after so many bits. To overcome this each character is suffixed with one or two (according to how the devices are set) 'stop' bits, which are always low. After each character we always have a low-to-high transition which can be relied upon to keep the two devices in step.

Notice that there is nothing particularly special about the start and stop bits. They look like any other bits, except that there is always a low-to-high transition at least once per character, and it's between these two bits. I point this out because it means, for example, that if you are sending serial data to a printer, if the signal is momentarily disconnected (transmission continuing but reception interrupted) then upon reconnection the printer will probably not be able to interpret the incoming stream of highs and lows. The printer will be confused until the next pause in transmission, unless the combination of received characters enables the printer to determine where the stop-start location is.

Transmission speed: 'Baud Rate'

Naturally, both sender and receiver must be set to the same nominal communications speed. This speed is measured in bits-per-second, a unit also known as the baud. (One bit per second is one baud.) Commonly used rates are: 110 and 133 (for Selectric terminals, for example), 300 baud (modems communicating via telephone), 600, 1200, 2400, 4800 and 9600 baud. Some devices also communicate at 19 200, 38 400 and even 76 800 baud, but such are rare.

Parity

An embellishment which is occasionally seen is the use of 'parity' as an error checking method. In a seven-bit code, for example, an extra bit may be added after the last bit (but before the stop bit). The sender counts the number of 'one' bits in the character, and if the answer is even it sets the parity bit to 'one', if not it is made 'zero'. (This is the even parity convention. There's an equally little-used odd parity convention which makes the parity bit 'one' for an odd total.)

When the receiver gets the character it does the same arithmetic and compares its answers to the parity bit received with the character. If it has the same answer it knows all is well; if the answer is wrong an error has occurred somewhere. For example, suppose an 'A' is transmitted (seven-bit, even parity). This would be represented as in Figure 3.

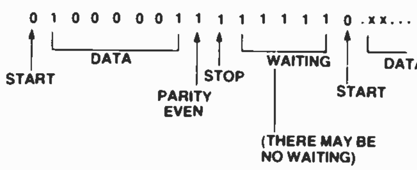


Figure 3. **Representation of the letter 'A' in seven-bit even parity code. Note that a '1' is a low voltage and a '0' is a high. The line normally sits at low or '1'.**

Now if one of those bits were accidentally changed somewhere along the way, there would be either one or three '1's, which is an odd number and does not agree with the parity bit. (And of course if the parity bit was accidentally changed, it wouldn't agree properly either.) You can probably see that this scheme cannot show where the error occurred or how to fix it, nor does it signal double errors. It is basically a low-overhead warning device.

In fact parity is generally ignored, since most personal computer equipment is not operated in electrically noisy environments where such errors are likely to occur, and in any case such equipment has no convention for requesting that the sender resend the faultily received data. (Often the receiving device may be set to expect the parity bit but not use it.) However, I have included this description so that you know what parity is when the equipment has a switch to select or deselect its use.

Lots of options!

As you can see, even thus far there are plenty of options to choose from. In a typical device many of these options may be switch selectable, usually miniature DIP switches inside the box, or perhaps soldered jumpers. In some cases, such as terminals and computers, some of these features may be programmed from the keyboard or from software.

So there are plenty of ways in which your two little darlings won't be able to talk to each other! But wait, there's much more!

How many duplexes?

Although not strictly of direct concern in the RS232 interface, some equipment, particularly terminals and modems, provide a 'Full/Half Duplex' switch.

'Full Duplex' means that when the terminal transmits a character to the remote computer the computer immediately echoes the character back to the terminal, whereupon it appears on the terminal's screen (or paper, if a teletype). If there is no echo then the character you typed will not appear on the terminal's screen. This is a kind of insurance method to let you know that the computer is listening.

In 'Half Duplex' set-ups it is assumed that the computer will not echo the characters from the terminal, and thus the terminal puts the typed characters on the screen whether or not the computer is awake.

The surprise comes if you have your terminal (or modem) set to Half Duplex, *and* the computer you are talking to echoes the characters. Then if you type 'FRED' you'll see 'FFRREEDD'.

Not so fast!

A commonly needed feature is the ability to tell the sending device to slow down. I don't mean to send at a lower baud rate, but rather to pause for a moment. A typical situation where this occurs is in slow printers. When the carriage reaches the end of the line the printer must tell the sender to wait until the carriage returns before sending more characters.

Such a signalling system is known as 'handshaking'. Typically this is implemented by adding an extra wire to the interface cable. The receiver maintains this wire at a 'high' signal level while it's OK for the sender to send, pulling it 'low' to tell the sender to halt the flow of data. Sometimes an interface will have handshaking lines both ways, so that either device can halt the other.

A complete two-way interface would consist of two data wires, two handshaking wires and ground — a total of five ▶

wires. Most RS232 hook-up problems occur because one piece of equipment needs some of these signals which the other does not provide, or because the wires in each piece of equipment are not connected to the corresponding pins in the interfacing connectors.

Not so fast type two

A quick note here that on some intelligent printers handshaking is carried out using a method called 'X-on, X-off'. Instead of a separate handshaking wire, the printer has a data output wire (normally printers only receive data). If the printer wishes to halt the sender the printer sends a control character to the sender (usually control-S, hex 13, which is also known as 'Direct Control 3'). Subsequently sending the same character will restart the data. Note that this is the same character which you use in CP/M (and Apple) to stop and start a continuous display to the screen from the keyboard.

Handshaking and buffers, etc

How necessary is handshaking in practice? A major sore point in the small computer industry has been the need for handshaking in printers. The Epson MX-80, for example, was available at one time with a serial interface known as 8141. This interface could only remember a maximum of two characters as they arrived from the computer. Since the 'line-feed' time exceeds the time of two characters, even at the slowest baud rates it was necessary for the interface to signal a halt after each line. The Exidy Sorcerer and the standard Apple printer interface board do not have any handshaking inputs, and consequently it would be impossible to make this combination of equipment work serially. (This particular problem rarely comes up since the MX-80 has a parallel input which is usually used. The Sorcerer has a parallel output, and the Apple has available for it a parallel printer board. I am simply showing how close to the surface such problems are swimming.)

A solution to this dilemma which is finding widespread adoption is to incorporate a 'buffer' into the serial interface. Such is the Epson 8145 interface, which has a 2000-character (approx.) buffer. Since the MX-80 chugs along at 80 characters-per-second (cps), if the computer transmits at 300 baud (30 cps) the buffer is normally virtually empty. At line-feed time the buffer fills up a little as the computer continues to transmit. But the MX-80 catches up on the next line. There is thus no need for handshaking. You can, however, get into trouble if the computer sends a large number of form-feeds, which take a long time.

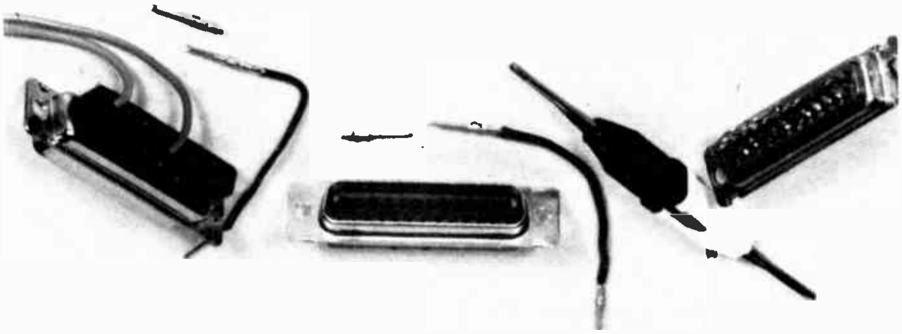


Figure 4. **Photo showing two different styles of RS232 connectors. In each case the individual wires are soldered to the connector pins or receptacles.**

To the left is a connector which comes with all the pins or receptacles permanently fixed in place; wires are soldered into 'cups' on the rear, which is the side in view here.

In the centre and to the right are shells which come 'empty', into which may be inserted male pins (into centre shell) or female receptacles (right). These can be more convenient, as the pins (shown separately and attached to wires) are easier to wire to before being placed in the shell, and in fact when installed are well separated by the shell (as can be seen in the rear view of the female connector on the right). The tool on the left is for the installation or removal of the male and female contacts.

In each case the connector may be bolted to a chassis, or put in a plastic cover for use as on the end of a cable.

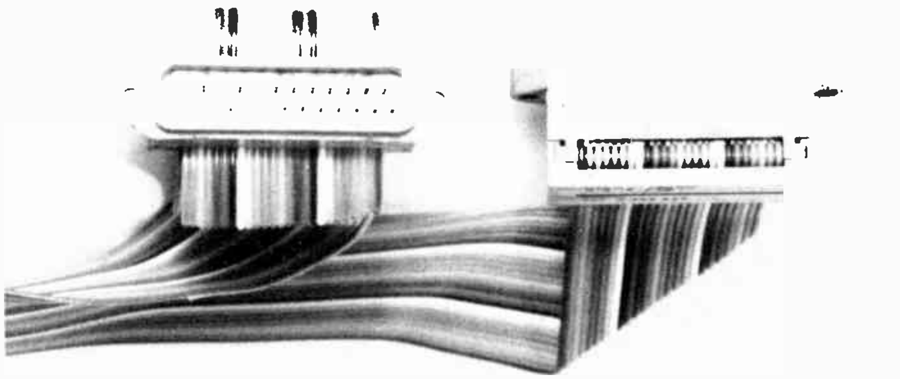


Figure 5. **These are called 'Insulation Displacement Connectors' (IDC) and of course must be used with ribbon cable. They can only be used if it is desired to connect all 25 pins at one end to all 25 at the other. However, they are very easy to install; all that is needed is a small vice to squash the connector onto the ribbon.**

Wires and connectors and stuff

The connector used with RS232 is known as a 'DB25', which has 25 pins in the male, and 25 receptacles in the female. Various styles are shown in Figures 4 and 5, with pin numbering shown in Figure 6. But why 25 pins?

RS232 was endowed with a pile of features not now used, and these were implemented using most of the 25 pins. Now very few of the pins are used. The extra pins provide two opportunities for confusion and problems, however. One problem is that with such a profusion of pins it can be difficult to figure out which ones you are supposed to use for your application.

'Business end' of male, or solder side of female.

1 2 3 4 5 6 7 8 9 10 11 12 13
14 15 16 17 18 19 20 21 22 23 24 25

'Business end' of female, solder side of male.

13 12 11 10 9 8 7 6 5 4 3 2 1
25 24 23 22 21 20 19 18 17 16 15 14

IMPORTANT NOTE: This numbering scheme means that with the IDC connectors the pin numbers do *not* correspond to the ribbon conductor numbers; 1 will be 1, but pin 14 will be ribbon conductor 2, etc.

Figure 6. **DB25 contact numbering.**

The second problem area is that with all those extra tantalising pins available and otherwise doing nothing, many manufacturers use the 'spare' pins for other purposes. Exidy uses them for the cassette interface. IDS, in their Paper Tiger printers, use the same DB25 for both serial and parallel interfaces.

'Official' Signal Name	Abbrev(1)	Pin No.	DTE 'Terminal'	DCE 'Modem'	Comments
Protective ground	PG	1	—	—	Optional
Signal ground	SG	7	—	—	Necessary
Data:					
Transmitted data	TxD	2	Out	In	
Received data	RxD	3	In	Out	
Handshaking:					
Request to send	RTS	4	Out	In)	Basically
Data terminal ready	DTR	20	Out	In)	same use
Clear to send	CTS	5	In	Out)	Basically
Data set ready	DSR	6	In	Out)	same use
Connector Sex:			Male(2)	Female	

(1) Note that the handshaking lines are sometimes indicated as inverted signals (e.g: $\overline{DTR}$). The idea is that if for the data a low is a '1', then if the data terminal is ready it should send out a '1'. In fact it sends out a high, which corresponds to a zero, hence the desire to use inverted signal notation. This refers, however, to the identical signal. In contrast there is the rare occasion when the equipment actually does put out an inverted signal, i.e: low means ready, high means not ready. Yeah, I know, but don't complain to me!

(2) In fact almost all terminals use female chassis-mount connectors. (A notable exception is the Heathkit H19.) It seems that it is almost standard practice to use females on equipment chassis, and male on cables (except for much DEC equipment, which uses male chassis mounts on equipment, and female connectors on cables). Note that this means you can't tell the DTE/DCE gender from the sex of the connectors.

Figure 7. **Table of signals, what they do, and connector pin assignments.**

That's fine except that if between such units you use a cable with too *many* wires implemented (and this can easily be the case if you use a standard RS232 cable in a set-up which does not use handshaking) then you are likely to blow something at one or both ends!

The pins which *are* commonly used are shown in Figure 7. Note that the naming convention can result in a variety of confusions. If the equipment is masquerading as a DCE the manual may tell you that, for example, pin 2 is 'Transmitted Data', which strictly speaking is an *input*. However, the manual writer may not know this and instead call it 'Received Data', intending 'Received' in a looser sense.

Fighting back

The first thing to do before connecting *anything* is to make yourself a chart like the one in Figure 8 for each piece of equipment you may have to connect together. This is *especially* important if you are involved with many different units. I have a whole binder full of such charts on the equipment I work with. Using this binder I can almost instantly connect any two units with few problems.

The point to this chart is that for each of your pieces of equipment (and I assume you're working with at least two!) it serves to collect the titbits of information you will glean from the manuals, the schematic and so on. You end up with the info in the same format for each unit, where it can be simply compared to give you the best idea of how to wire things up *before* you blow anything, and before you have the frustrating experience of having the system not work.

If handshaking lines are provided, try to find out if they actually do anything, or if they are dummies. For example, one printer may have an output which signals the sending computer to halt. Another printer may claim to have the same handshaking output, but it is actually internally wired permanently high, and is provided merely for supposed compatibility to a computer which may need such an input so as not to halt. Got that?!

Wiring up the cable

You will notice that if one of your units is a true DTE and the other a true DCE then a standard cable (pin 1 goes to pin 1, 2 to 2 and so forth, which is called a 'straight-through') will work. You are unlikely to see this situation very often, which is something you should know before you buy such a cable made up (they're likely to be expensive ready-made), or before you get convinced by the salesman that the printer hook-up is trivial.

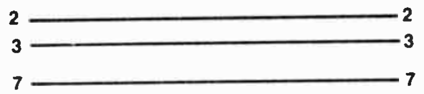
So you decide to wire your own cable. First, of course, you must obtain the appropriate sexes of connectors to mate with what you have on the equipment, and a cable with a sufficient number of conductors. If it's over 20 feet you may wish to use shielded cable, but I've used unshielded up to several hundred feet. ▶

PIN NO.	SIGNAL ABBREV.	SIGNAL NAME	IN/ OUT	OPEN OK?	COMMENTS
1					
2					
3					
4					
5					
6					
7					
8					
9					
10					
11					
12					
13					
14					
15					
16					
17					
18					
19					
20					
21					
22					
23					
24					
25					

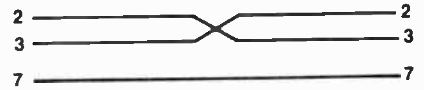
EQUIPMENT TYPE: CONNECTOR SEX ON EQUIPMENT: DATE:

Figure 8. **Interface chart to save you headaches.**

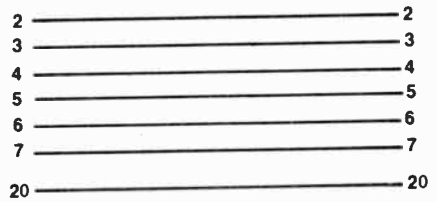
Notes



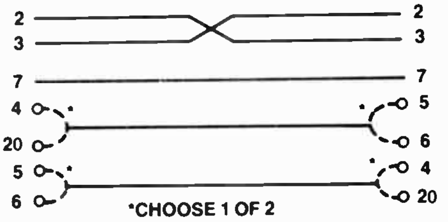
9a. The simplest possible cable, opposite sexes and no handshaking.



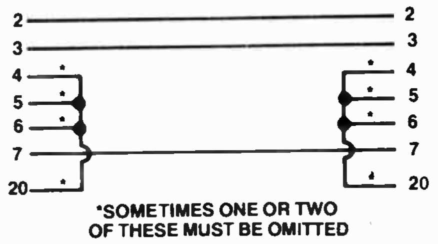
9b. The next simplest cable, DCE to DCE or DTE to DTE, and still no handshaking.



9c. Cable to join opposite sexes with handshaking. ('Standard' straight-through cable.)



9d. Joining same sexes with handshaking (one possibility).



9e. Joining opposite sexes, with defeated handshaking at both ends.

Figure 9. Some typical cable hookups.

Next, no matter what the equipment involved, wire pin 7 to pin 7. If it's a straight-through you are making, then go right ahead, 2 to 2, 3 to 3, etc.

The next-most-delightful situation is where the two units are of the same sex and need no handshaking lines. For the data lines simply wire 2 to 3 and 3 to 2.

If handshaking lines *are* needed then determine which handshaking outputs actually mean something (as opposed to the dummies). Then connect these to the handshaking inputs of the opposite units.

You may have a sender which is sending to a receiver which does not need to halt the sender. If this is the case you need to decide what to do with the sender's handshaking input. In some units it can merely be left open (unconnected), and this is seen as the same as 'high'. On other units open is taken as a 'low' and halts transmission. The handshaking input may be wired permanently high by jumpering it to a handshaking *output* on the *same device*. This is normally done inside the plug on that unit's end of the cable. Figure 9 shows some typical cable configurations.

The initial hook-up

Armed with the appropriate (we hope) cable, plug in and see if it works! It probably won't, so refer to Figure 10, which is a summary of all the things to check to make the two pieces of equipment compatible communicators.

A test box to defeat all problems

So perplexing are some RS232 problems which I have encountered that I highly recommend obtaining a test unit of some kind if you are going to be involved with many such situations. In the next part of this article we intend to present details of a device which is designed to handle these problems, and which also permits quickly patching together any trial interface configuration. Unlike commercial units, it will even enable you to determine the inputs and outputs of a completely unknown interface with no documentation. ●

1. Number of **bits per character:** 5, 6, 7, or 8.
2. Number of **Stop bits:** 1 or 2.
3. **Baud Rate:** 110, 150, 300, 600, 1200, 1800, 2400, 4800, 9600 or other.
4. What to do with **Parity:**

   On transmission: No Parity, Even Parity, Odd Parity, Parity bit set to 0, or Parity bit set to 1.
   On reception: No Parity expected, Ignore Parity, Expect Odd, or Expect Even.

5. **Full** or **Half Duplex.**
6. Make sure machines are **On Line** if they have the ability to be off line.
7. A rather rare final item which can cause problems is an option on a few machines which allows for the inversion of the polarity of the data signals and/or handshaking signals. You should set these to: Negative Mark for the data lines, and handshaking lines should indicate OK to proceed with a high level, STOP with a low level.

Figure 10. List of quick checks to make when hooking up two pieces of gear for the first time.

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RS232 serial interface trouble-shooter

Making a 'standard' RS232 interface work can be a nightmare. 'Standards' not withstanding, you can regain lost sleep with this trouble-shooting unit.

Grahame Wideman

THE PREVIOUS feature explained how computer serial interfaces of the RS232 type are supposed to work, and why they frequently won't. Here we present the design and construction details of a test unit which solves most of these problems. You may wish to build it as described here, or simply borrow the principles to troubleshoot interfaces using other instruments.

The troubleshooter provides the capability to patch together any wiring arrangement, and to monitor what is happening on each wire. In this much it parallels the better commercially-available RS232 'problem solvers'.

However, it also includes an apparatus for determining exactly which interface wires are outputs, inputs, not connected, or shorted, thereby making possible a complete picture of a totally unknown interface. This is extremely useful if the equipment in question has no manual, or as is more likely, has a manual which leaves the subject of the RS232 interface completely ambiguous.

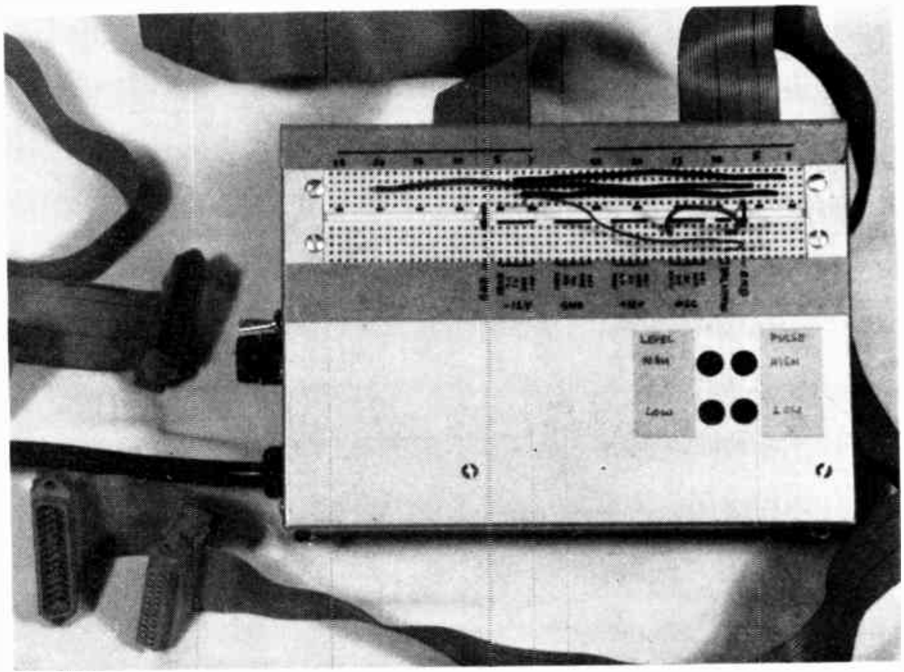


Figure 1. **The RS232 troubleshooter.**

Patching board

The heart of the troubleshooter is a breadboard patching block which is wired permanently to a pair of ribbon cables, each cable having attached to it both male and female DB25 connectors of the 'insulation displacement' squash-on variety. After peeling the adhesive plastic backing off the breadboard block (and cleaning it up a bit), the individual conductors of the ribbon cable are soldered to the *underside* of the rows of contacts in the breadboard, as shown in the photo of Figure 2, and detailed in the diagram of Figure 3.

This simple device already gives two capabilities, as shown in Figure 4. First, both cables can be attached, one to each of the pieces of equipment which are to be interfaced together. Having both a male and female connector on each cable ensures that plugging in will be no problem. Then the particular pin-to-pin wiring can quickly be tried out using jumper wires on the tester's breadboard patching area, before a permanent cable is made up.

The second way to use the device as so far described, is for 'tapping into' an existing interface arrangement which is now perhaps malfunctioning. Suppose

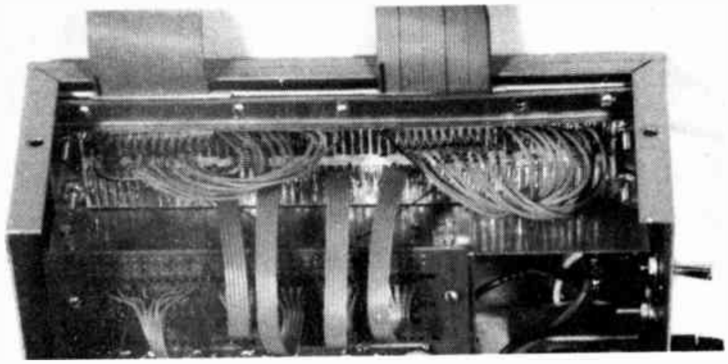


Figure 2. **Closer view of breadboard area used for patching the troubleshooter's two DB25-equipped ribbon cables, and for connection to the unit's signal monitor and test signals.**

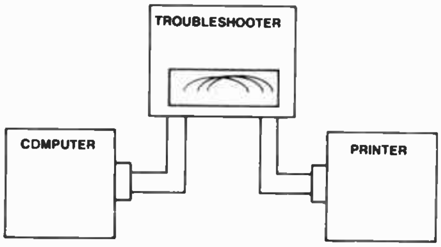


Figure 4a. **Using the troubleshooter's patching area to rig new trial cable before making permanent version.**

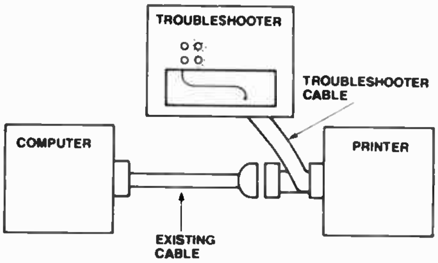


Figure 4b. **Using just one ribbon cable, the troubleshooter provides a convenient way to tap into and monitor the signals on a 'supposed to be working' RS232 hookup.**

the 'existing system' is a computer talking to a printer. Simply unplug one end of the computer-to-printer cable (let's say the printer end), and plug it into one of the two connectors (male or female as appropriate) on one of the troubleshooter's ribbons. Plug the remaining connector on the *same* ribbon into the receptacle on the printer. Now the short length of ribbon between male and female DB25s carries the connection from computer to printer, and in addition this ribbon brings out all 25 lines to the breadboard where they may be conveniently monitored.

Signal monitor

In order to monitor the signals on an RS232 line it is possible to get away with simply a LED with a resistor in series. However this loads the line, possibly changing the conditions you were trying to monitor. Additionally, you would not see any quick pulses of activity which may be important.

Consequently the monitor incorporated in this tester has been designed to address these two problems. Four LEDs are used, two to indicate a steady high or low level, while the other two flash on for about a half second in response to a positive or negative pulse. The level LEDs respond only to valid high or low signals; a voltage in the middle (around zero), or an open line will cause neither LED to illuminate. The LEDs are powered from a built-in mains power supply, and the RS232 line is monitored via high impedance buffers, so as not to disturb it.

(It should be remembered here that the RS232 line levels are −3 V to −12 V for a 'low' representing a data '1', and +3 V to +12 V for a 'high' representing a 'zero'. With no data the line sits at low.)

The monitor input is soldered to the underside of a contact strip on the breadboard (actually two strips, in case of wear), and thus may be patched to any other contact as desired for observation of the signals there.

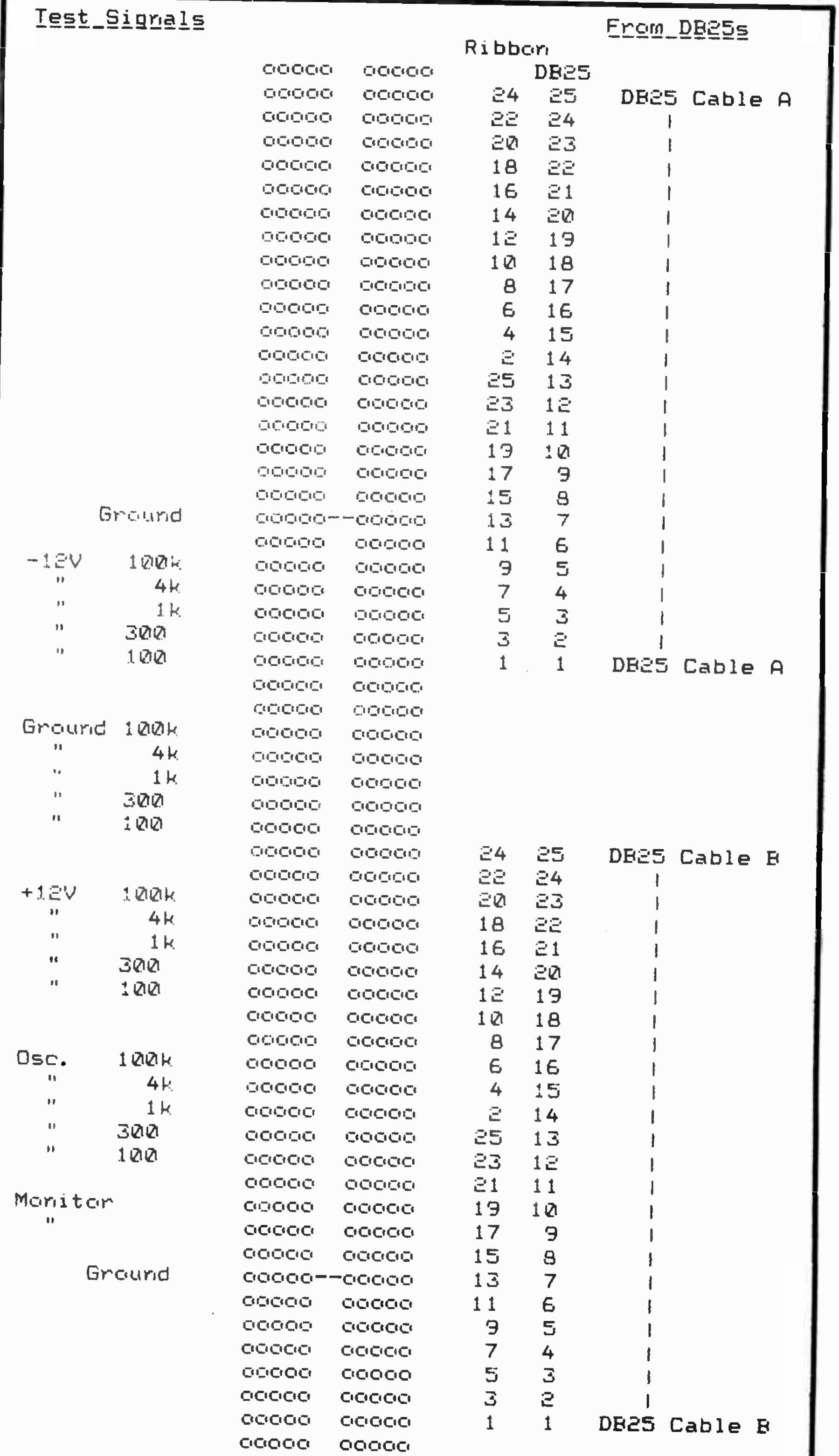


Figure 3. **Diagram showing one possible arrangement for the breadboard connection area. Note that the DB25 ribbon cables occupy one 'side' of the board, and the test monitor and signals occupy the other. Opposite each ribbon's pin number 7 is a ground connection, across which would normally be installed a small jumper.**

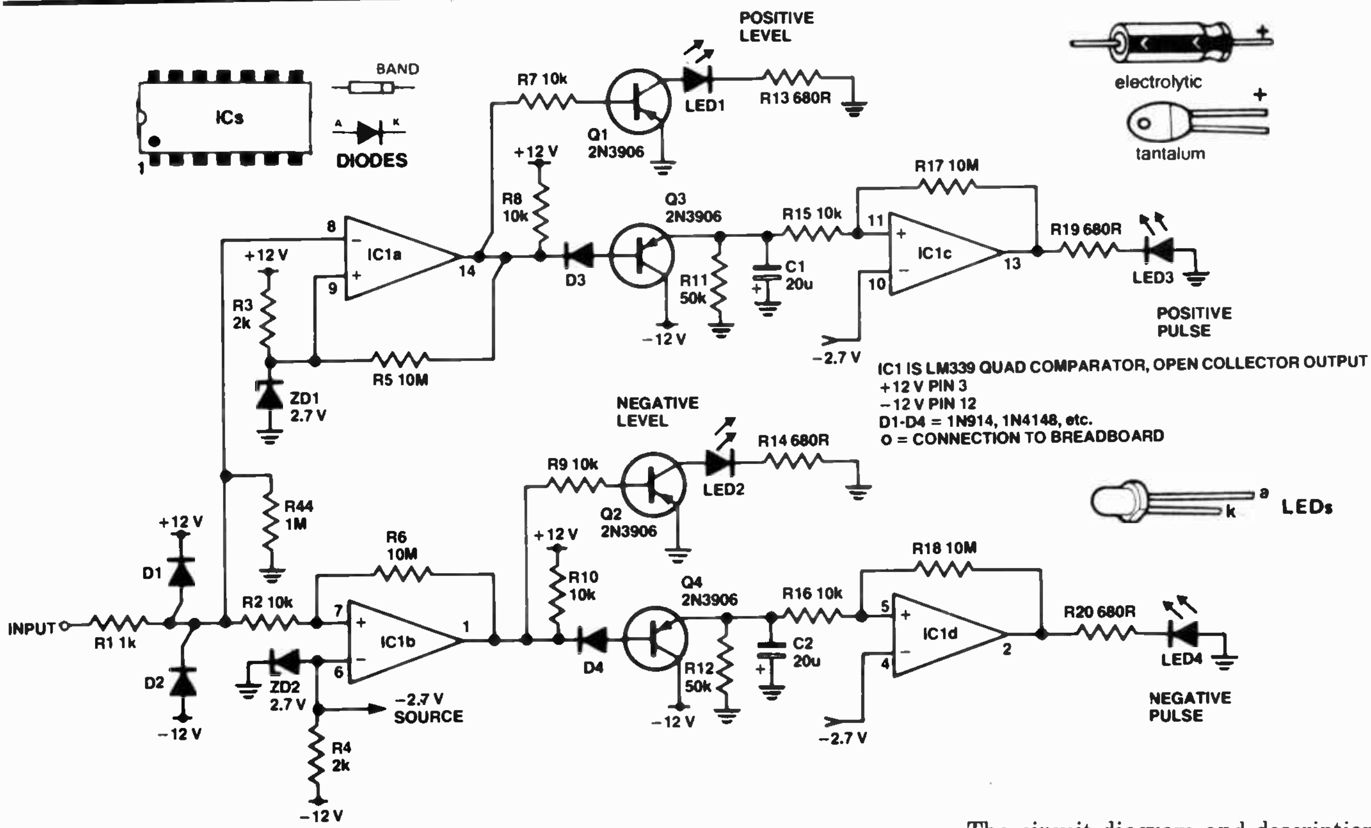


Figure 5. **Circuit diagram of monitor section.**

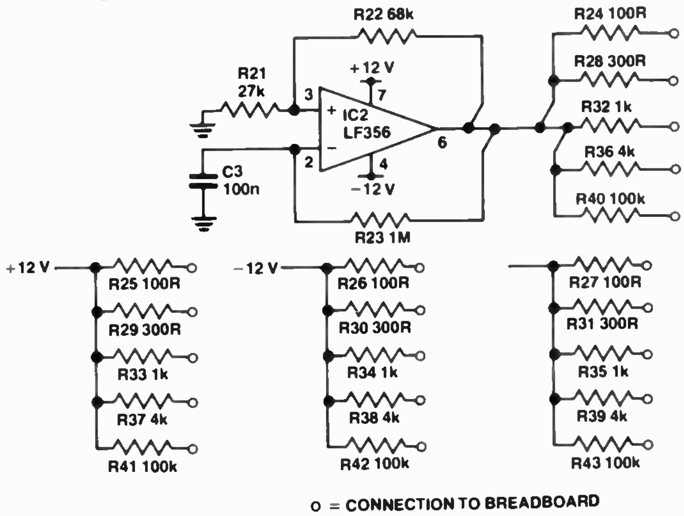
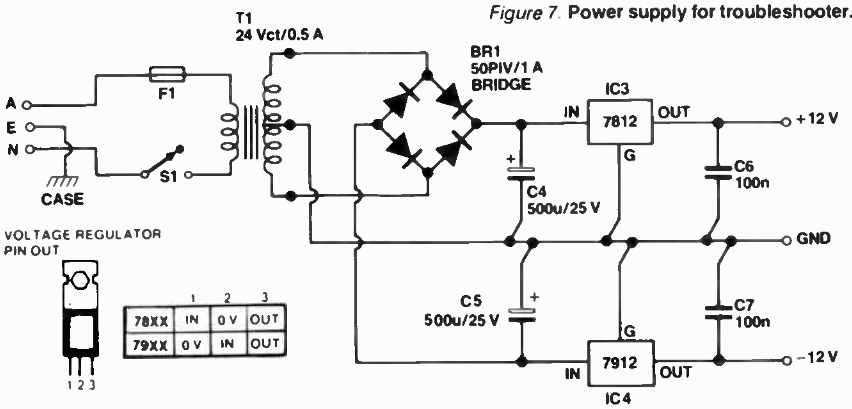


Figure 6. **Circuit of test oscillator, and other test levels.**

Figure 7. **Power supply for troubleshooter.**



	1	2	3
78XX	IN	0 V	OUT
79XX	0 V	IN	OUT

The circuit diagram and description for the monitor, the test oscillator to be described, along with the power supply, are to be found in Figures 5 to 7.

Generated signals

Of immediately obvious use are the 'high' and 'low' signals provided. These may be used to apply 'halt' or 'go' signals to handshaking lines. Additionally there is built-in a square-wave generator which continuously oscillates between high and low conditions at a rate of approximately five times per second.

Each of these signals, along with ground, is supplied to contacts on the breadboard via a selection of resistors, from 100 ohms to 100k ohms. The usefulness of this arrangement may not be immediately apparent, and for explanation I must describe the electronic circuits which transmit or receive on an RS232 line.

How it works

The monitor

The schematics for the troubleshooter's main monitor, test oscillator and power supply are to be seen in Figures 5 to 7.

Components R1-D1-D2 prevent the input signal from causing damage should it happen to exceed the troubleshooter's power rail limits. From there the input signal is routed to two very similar 'channels', one concerned with 'high' levels and pulses, the other with 'low' levels and pulses.

Looking then at the positive channel, the input signal arrives at the negative input of comparator IC1a, where it is compared to a reference of about 2V7 (which is set by ZD1 at the positive input of IC1a). Supposing that the input signal exceeds 2V7, then IC1a's output is low, turning on Q1 via R7, and illuminating LED1 to indicate a 'high level'.

At the same time, the low level (about −10 V) at the output of IC1a turns on Q3, quickly charging C1 'down' to about −9 V. IC1c sees this voltage and compares it to −2V7, sees that it is lower and lowers its own output illuminating LED3 to indicate a 'high pulse'.

If the monitor input now drops below 2V7, IC1a's output will go high, turning off Q1 and the 'high-level' LED1, and also Q3. However the 'high pulse' LED3 will remain on for a short while (about a half second) as C1 is charged up past the 2V7 point by R11. Notice that this delayed LED3 action would have occured even if LED1 had been on for only an invisibly short length of time. Hence LED3 makes visible short pulses which cannot be seen by simply watching the level, whether on the troubleshooter's level LEDs, or even with an oscilloscope.

The negative channel works similarly, the only change being to swap the positive and negative inputs of the input comparator.

Test oscillator

As we shall see, IC2's output must sit in either high (about +10 V) or low (about −10 V) states. Let us assume it is initially low, and that C3 starts out uncharged, so that there is 0 V at the op-amp's negative input.

Since the op-amp output is at say −10 V, the positive input will be at approximately −3 V, established by the R21-R22 voltage divider. Remembering that we assumed the negative input to be at 0 V, the 'low' output will remain temporarily unchanged.

However, the low output will charge C3 via R23 downwards. After a while the op-amp negative input will be less than its positive input (at −3 V), and thus the output will change states to +10 V. When this happens the voltage at the positive input changes, of course, to +3 V, maintaining this state of affairs.

Again we must wait for C3 to be charged via R23, this time up to +3 V. You should be able to see that this oscillating action will continue, and that the period is the time taken for C3 to charge from −3 V to +3 V, then back down to −3 V, when R23 is pulled up to +10 V and then down to −10 V respectively.

The values given provide a frequency of about 5 Hz, quite suitable for this testing purpose.

The oscillator output is delivered to the breadboard area via various values of resistor, as described in the text.

Of drivers and receivers

For various reasons, special purpose buffers are used to send signals and receive signals on an RS232 line. These are called 'drivers' and 'receivers', and are exemplified by the National LM1488 and 1489 respectively. Figures 8 and 9 show a simplified view of how the driver output and receiver input look electrically.

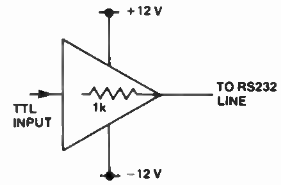


Figure 8. **Simplified view of an RS232 'driver' output.**

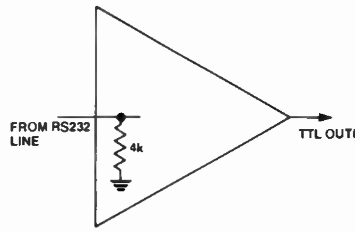


Figure 9. **An RS232 'receiver' looks like this electrically.**

The points to notice are that a receiver input looks like a (approximately) 4k resistor to ground. An operating driver output looks like a (approximately) 500 ohm resistor pulling up to 12 V (nominally), or pulling down to −12 V, according to its state.

Knowing these facts it is apparent that if a high or low signal is applied through a resistor to a receiver input or driver output, the resulting signal on the line (which can be monitored) will be high, low or in-between depending upon the value of resistor used.

Therefore, when looking at an unknown line, by applying the test oscillator's output via each resistor in turn, it is quickly possible to tell what that line does. The chart in Figure 10 details this.

cont. on page 69

Parts List

Resistors .... ¼ W unless specified

R1	1k
R2	10k
R3, 4	2k
R5, 6, 17, 18	10M
R7, 8, 9, 10, 15, 16	10k
R11, 12	50k
R13, 14, 19, 20	680
R21	27k
R22	68k
R23	1M
R24, 25, 26, 27	100
R28, 29, 30, 31	300
R32, 33, 34, 35	1k
R36, 37, 38, 39	4k
R40, 41, 42, 43	100k
R44	1M

Capacitors

C1, 2	20u/20 V electrolytic
C3, 6, 7	0u1 tantalums
C4, 5	500u/25 V electrolytic

Diodes

D1, 2, 3, 4	1N914 or 1N4148 etc
ZD1, 2	2V7 small zener diode
BR1	Bridge rectifier, 50 PIV/1 A
LED1, 2, 3, 4	LEDs of your choice of colour

Transistors

Q1, 2, 3, 4	2N3906

Integrated circuits

IC1	LM339
IC2	LF356A
IC3	7812
IC4	7912

Transformer

T1	240 V primary, 12-0-12/500 mA secondary

Miscellaneous

Breadboard, case, fuse and holder, switch for power, power lead and plug etc.

Line Condition	Oscillator Signal via Resistor (ohms) 100k	4k	1k	300	100
Open	HL	HL	HL	HL	HL
Receiver Input	None	HL	HL	HL	HL
Driver Out-Low	L	L	L	FL/HL	HL
Driver Out-High	H	H	H	FH/HL	HL
Short to Ground	None	None	None	None	None
Short to +12 V	H	H	H	H	H
Short to −12 V	L	L	L	L	L

Charts shows the Level LEDs (not 'pulse' LEDs) activated in various cases.

H = High; L = Low; None = neither LED on; HL = alternating H and L; FL = Flashing Low; FH = Flashing High

Figure 10. **Chart showing how to test an unknown RS232 line, and the monitor's indications under various conditions.**

Why some CMOS circuits don't work as you expect

Stephen Dolding

A 4093 is a 4093, right? Well . . . yes, and no. There are quite a few pitfalls in the CMOS 'jungle' and it's handy to know about them before venturing forth.

CMOS '4000' SERIES integrated circuits are manufactured by at least six major manufacturers and the 74C series by at least two major manufacturers, but it must not be assumed that a 4XXX from one manufacturer is interchangeable with a 4XXX device from another manufacturer. This article explains some of these differences.

Schmitt gate oscillator

What could be simpler than the oscillator circuits shown in Figures 1(a) and 1(b)? There are so few components that you would expect these circuits to work first time.

You have selected stable, close tolerance components and calculated the frequency according to the formula:

$$f_{out} = \frac{1}{CR\log_e\left[\left(\frac{V_{cc}-V_{t-}}{V_{cc}-V_{t+}}\right)\left(\frac{V_{t+}}{V_{t-}}\right)\right]}$$

where C = capacitor value in uF
R = resistor value in kilohms
f_{out} = frequency out in kHz
V_{cc} = supply voltage
V_{t+} = upper trigger level of Schmitt trigger
V_{t-} = lower trigger level of Schmitt trigger

Now, the question is, "what are the values of V_{t+} and V_{t-}?" We need to refer to the manufacturer's data sheet for an answer. But which manufacturer? There are at least six different manufacturers to choose from and each one gives a different range of possible values for V_{t+} and V_{t-}.

Considering the 4093 IC (Figure 1(b)); if all the databooks are consulted it is found that for V_{cc} = +5 V the *highest* typical value of V_{t+} = 3.6 V (extreme = 4.3 V).

The *lowest* typical value of V_{t+} = 2.7 V (extreme = 1.7 V).

The *highest* typical value of V_{t-} = 2.2 V (extreme = 3.3 V).

The *lowest* typical value of V_{t-} = 1.4 V (extreme = 0.7 V.

Now if V_{t+} = 3.6 V, V_{t-} = 1.4 V, V_{cc} = +5 V, C = 100 nF and R = 10k, then from the formula above f_{out} = 52.9 Hz.

With the same type from another manufacturer, V_{t+} could be 2.7 V and V_{t-} could be 2.2 V, in which case recalculation gives a frequency of 249 Hz!

These figures are based on typical values of trigger level. Extremes of high and low trigger levels could give frequencies ranging from 27.5 Hz to 961 Hz with the same values of C = 100 nF and R = 10k. This gives a frequency range of almost 35:1 if the whole spectrum of possibilities is considered.

Now it will be clear why the circuit may oscillate at a frequency which is considerably different from what was expected or intended by the designer who only consulted one manufacturer's databook!

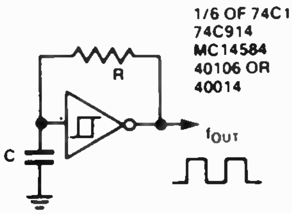


Figure 1a. Different manufacturers' ICs will produce different results.

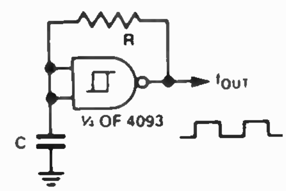


Figure 1b. Such a simple circuit but the value of f_{out} can vary from 52.9 Hz to 249 Hz, depending on the brand of 4093 used.

The monostable

Now let us look at another CMOS circuit often used by the hobbyist — the 4528 dual monostable, shown in Figure 2.

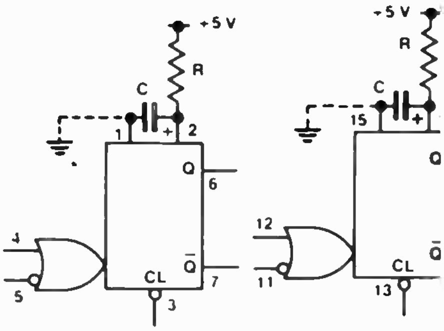


Figure 2. The 4528 dual monostable can have variations in its performance of up to ±50%.

Here again, different manufacturers give different formulae for the monostable time constant. A typical formula (for +5 V supply) is

$$t = 0.37CR$$

However, it could vary from t = 0.32CR to t = 0.42CR typical, with variations up to ±50%.

The pulse width depends very much on the supply voltage. Some manufacturers' 4528s give increasing pulse width with increasing supply voltage, others give a reverse effect.

The formula given above depends on the value of C being greater than 10n. For smaller values of capacitance the manufacturers data sheets need to be consulted.

It should also be noted that some manufacturers require pins 1 and 15 to be grounded externally for correct operation. To overcome the variations in timing formula, a CMOS 4538 integrated circuit can be used in place of the 4528. The 4538 is pin compatible with the 4528 IC and the formula is:

$$t = CR$$

with variations of only ±5%.

In all timing circuits using CMOS ICs it is wise to make provision for trimming the value of the timing resistor to allow for adjustment.

The counter/divider

Next we come to a well-known decade counter/divider IC — the 4017.

Will the circuit in Figure 3 always work correctly? No, only with a Motorola 14017 or an RCA 4017, because these have an internal Schmitt trigger on the clock input. Other manufacturers' 4017s do not, so false counting may result.

BCD decoder

Another fairly common integrated circuit likely to give problems is the 4028 BCD-to-10-line decoder. We now come to see that actual logical differences can occur between one manufacturer's 4XXX and another manufacturer's 4XXX.

With the 4028, some manufacturers (Motorola and RCA) do not decode the six 'illegal' binary codes 1010 to 1111 (i.e. 10-15), while other manufacturers (including National, Fairchild and Philips) decode these outputs as if the input was 8 (1000) or 9 (1001).

The problem of logical differences between one manufacturer's device and another manufacturer's device (with supposedly the same type number) applies also to the 4585 four-bit comparator and even to the ubiquitous 555 and 556 timers. There may be other examples too. The problem fortunately does not occur with the range of quad gates.

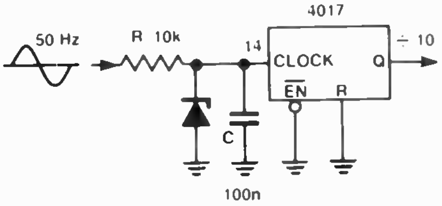


Figure 3. This circuit will only work correctly with particular brands of 4017 decade counter/dividers. The wrong choice can result in false counting.

The moral

The above-mentioned examples were all encountered during the design of one piece of industrial equipment which made use of these common CMOS parts.

You may well ask "If design engineers, who have ready access to all the data books, can run into such problems, what about the unsuspecting hobbyist, who has no data?"

The moral of this article is that "forewarned is forearmed". It is hoped that this article may at least prevent some construction projects from being abandoned because they do not appear to work correctly at first sight. Designers who publish projects should check that there are at least two manufacturers ICs which will work in the circuit as intended and, if necessary, spell out the names of suitable manufacturers in the parts list. Best of all, only design circuits that will work with all manufacturers devices of the same basic type number (though this may not always be possible).

If problems occur, all that may be required is to try an IC from a different manufacturer ●

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cont. from p67.

A word of warning is in order here however. Proper use of the monitor assumes that you at least know which pin on the connector is ground. This is almost always pin 7, so consistently in fact that we located the monitor's ground points on the breadboard opposite to the pins 7 of each of the two ribbon cables, and permanently left a small jumper installed at these two locations (see Figure 3). However, there are lurking about some units which don't abide by this standard. The only thing you can do about this (if you are documentationless and suspect this problem) is to open the case and actually trace the unit's circuit-board ground and see what DB25 pin it goes to.

Construction notes

The construction is not too critical. As can be seen the prototype version was built using Veroboard. One plan which is extremely useful to follow is to make a simple frame assembly like the one shown in Figure 11, which serves two purposes.

First it includes a flat clamp to grip the two ribbon cables (liberal use of double-sided wall-tile sponge adhesive tape also helps). Secondly, it keeps together the circuit board and the breadboard. In both these respects it makes wiring to the breadboard easier, and virtually eliminates any problems of wire breakage when the various parts are moved about during construction or testing of the project. As the photos show, the entire guts of the prototype can be removed in one piece, connected to the case only by the leads to the PSU.

Another hint: DON'T FORGET when soldering the ribbons to the breadboard that the numbering of the DB25 pins is *not* the same sequence as the ribbon cable conductors. This is shown in Figure 3.

Improvements

Although our troubleshooter has proved tremendously useful, I cannot claim that our prototype is the last word. In fact, I feel that if done again we would add several extra LEDs as simple on-off high-low indicators along with the existing level plus pulse monitor, so as to keep an eye on several lines at once. You may wish to adopt this idea in your unit. ●

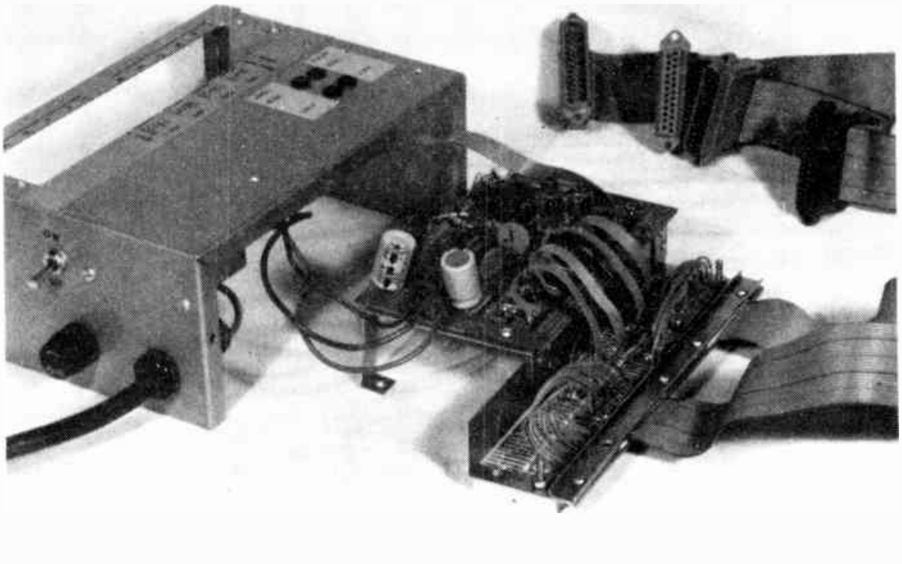


Figure 11. **When constructing a unit such as this, where there are a lot of wires hanging around, it's helpful to use brackets and cable clamps like the ones in this photo. They prevent undue strain on soldered connections, improving reliability, and enable the circuitry to come out of the box in one piece, more or less, for testing purposes during construction, or later if the unit needs repair.**

Principles and problems in loudspeaker design

This two-part feature describes in down-to-earth fashion, the principles and practice of loudspeaker design.

David Tilbrook

MORE MONEY can be saved by the construction of a pair of loudspeakers than any other single component of the hi-fi system. Unfortunately, they are also the most important hi-fi component! Unless the turntable or amplifier is particularly poor, the loudspeaker will undoubtedly determine the overall sound of the system. For this reason it is disappointing there are so few really good kit loudspeakers.

The fact that a "correct" loudspeaker doesn't exist is to be expected, since the principles of loudspeaker operation are enormously complex. Every loudspeaker model makes certain assumptions to simplify the mathematics and to make the model manageable. If these assumptions are overdone the model rapidly loses relevance, becoming incapable of making worthwhile predictions about the *real* loudspeaker. While it is true that a detailed understanding of loudspeaker operation is not necessary to enable a kit loudspeaker to be built, some understanding will enable the optimum to be obtained from the loudspeaker and is essential for those brave experimenters who would like to get involved in modifying the loudspeaker drivers. When designing a loudspeaker it is necessary to understand the mechanism of operation of the drivers. Only then can the best choice of driver, enclosure type and crossover be established. Although loudspeaker design is as much an art as it is a science, the loudspeaker that has been created with a motley assortment of drivers placed in a box with some "general purpose crossover" is more likely to sound like a dropped saucepan than a good loudspeaker!

The most common loudspeaker consists of several moving-coil direct-radiating drivers mounted in an enclosure. These cover different frequency bands within the audio spectrum. A crossover is used tc separate these frequency bands and feec them to the appropriate driver.

If the drivers used had perfectly fla frequency responses, were constan eight ohm loads with infinite powe handling, and the crossovers representec lossless transfer characteristics with nc untoward interactions with the drivers and if nature did not object to th reproduction of low frequencies ir confined volumes (i.e. if the speed o sound was one tenth the speed it is loudspeaker design would be a simpl matter.

Most of these problems can b summarised with one word . . . inertia This is that property of nature whereb things resist change. We can't really complain too strongly about inerti since it is responsible for much of the order that exists in the universe Nevertheless, in loudspeaker design i causes real problems. The signal voltage from the power amp, the magnetic fielc

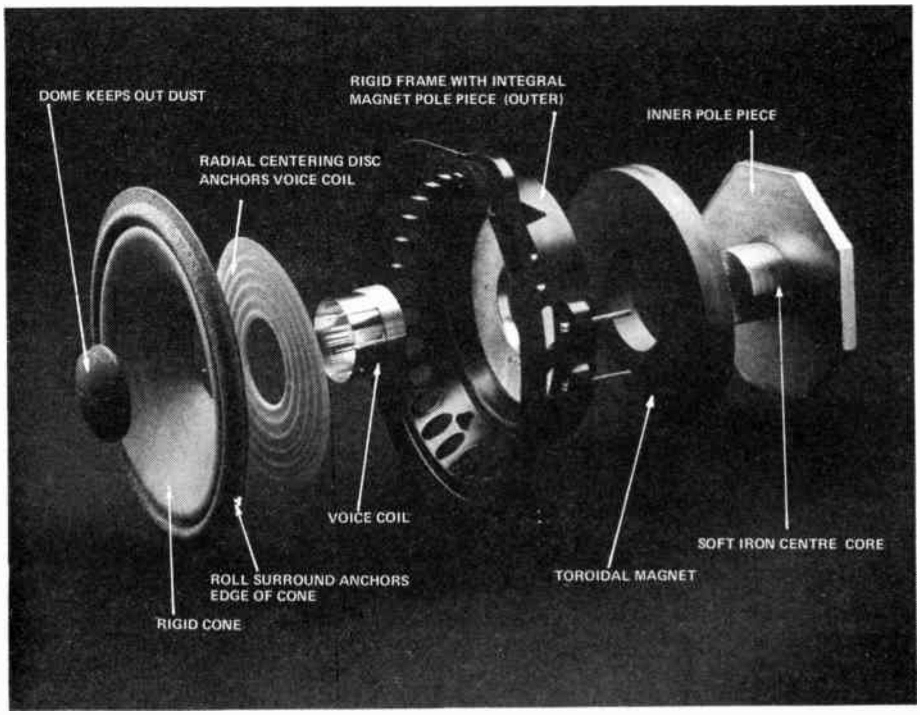
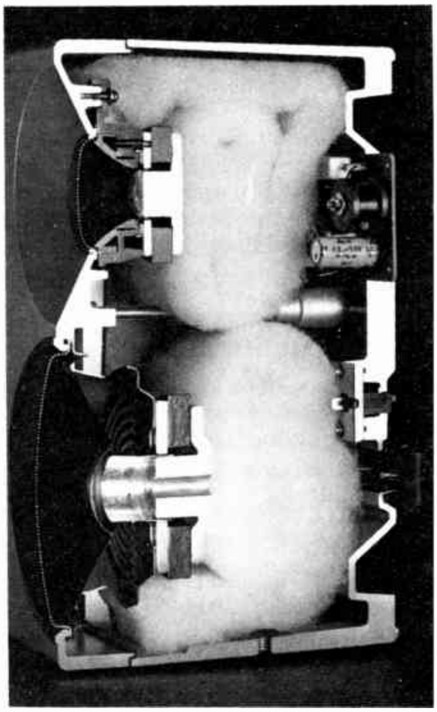


Figure 1. **An exploded view of a moving-coil loudspeaker showing the various components in its construction. Compare this with the cutaway view of a speaker at right (Pic: courtesy Bose).**



Cutaway view of a speaker unit showin internal construction (Pic: National).

around the voice coil, the movement of the coil and loudspeaker cone, all resist change. Since the objective of loudspeaker design is to convert an electrical signal into its exact acoustic counterpart, these sources of inertia cause errors resulting in distortion. The effects of inertia don't stop at just slowing down the system. The resistance to change of motion by the cone for example results in some parts of the cone moving before others. Sound waves start to travel along the cone itself, travelling radially out from the voice coil. Depending on the nature of the flexible surround between the cone and the chassis this sound wave will be partially reflected back down the cone. This causes constructive and destructive interference with the original sound wave propagating up the cone resulting in colouration. Clearly, this is not something the home constructor can do much about, since it depends on the manufacture of the particular driver concerned, but it indicates the sorts of problems that will be encountered.

The moving-coil direct-radiating speaker

The vast majority of drivers used in loudspeakers are of the moving coil type and as such all operate in a very similar way. Figure 1 shows a typical moving coil loudspeaker. Signal voltages from the power amp give rise to signal currents that flow through the voice coil. This is simply a coil of wire wound on a hollow circular former. In normal 8 ohm drivers the dc resistance of the voice coil is around 8 ohms, but the driver will only represent this resistance to the power amp at one specific frequency, the actual impedance of the driver varying widely as the frequency is varied (see Figure 2). A given signal voltage level will therefore produce different signal currents for different frequencies. The signal current causes a varying magnetic field to be produced around the voice coil. This field interacts with an intense magnetic field from the drivers' pole piece and magnet assembly causing a force to be exerted on the voice coil and loudspeaker cone.

As the cone moves it will compress or rarify the air immediately in front of it, creating an area of either increased or decreased pressure. These pressure variations comprise a sound wave that travels from the driver to our ears.

The electrical impedance of the driver is caused by several phenomena each one dominating in a specific frequency band. One of the most significant mechanisms is the back EMF (EMF stands for electromotive force, i.e: voltage) of the driver. The movement of the voice coil in the magnetic field acts as a generator causing a current to flow in the voice coil. This current is of opposite polarity to the applied signal current (another natural application of the principle of inertia) causing decreased current flow in the voice coil for a given signal voltage. This is seen by the amplifier as an increase in the drivers' impedance.

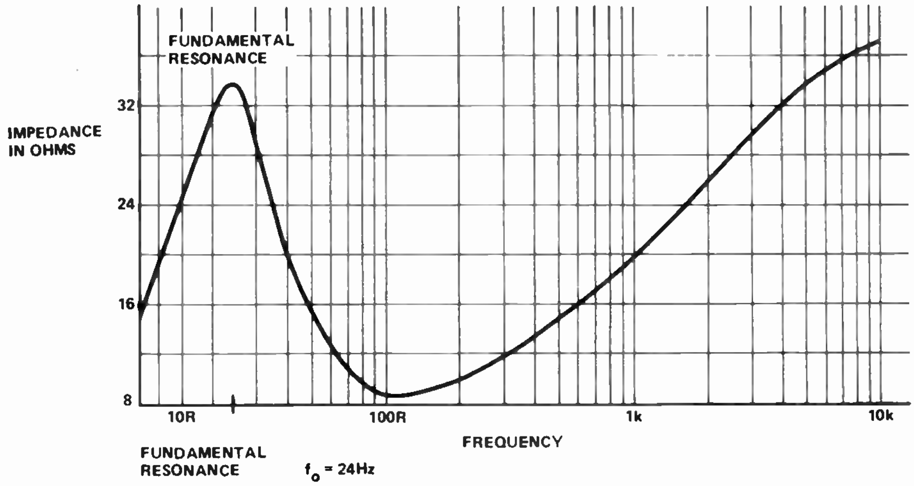


Figure 2. **Typical impedance versus frequency characteristics of a moving-coil loudspeaker**

EMF is given by the simple equation:

EQUATION 1

e = B1v where 'e' is the back EMF in volts
'B' is the magnetic flux
'1' is the length of wire in the magnetic field
and 'v' is the velocity of the cone

Since the magnetic flux and the length of wire in the magnetic field can be considered as constants, the equation shows that the amount of EMF generated is directly proportional to the velocity of the cone.

So the electrical impedance is a secondary phenomenon, is certainly not constant, and does not relate directly to the radiated acoustic power. The amount of back EMF will be determined by the velocity of the cone, and this is a function of nearly every major parameter of the loudspeaker box.

The force exerted by the voice coil on the loudspeaker cone is given by the equation:

EQUATION 2

F = Bil where 'F' = force on the voice coil
'B' = Magnetic field intensity
'i' = current in the voice coil
and '1' = length of wire in the field

Again, regarding 'B' and '1' as constants, the equation shows that it is current and not voltage that determines the force on the voice coil. Since the voltage contains the signal information from the power amp, it would be necessary for a perfectly linear relationship to exist between applied voltage and resulting signal current flow if a distortionless signal is to be produced. The impedance would have to be a constant and this is not the case. Fortunately the movement of the cone is not directly related to the current in the voice coil in the simple way shown above or the frequency response of a loudspeaker would simply be the inverse of its rather lumpy impedance curve.

In order to understand the parameters that determine the acoustic power actually radiated, it is necessary to look at the sources of mechanical rather than electrical impedance.

Converting energy

In the operation of a moving-coil direct-radiating driver there are really two energy conversions going on simultaneously. First the electrical energy is converted into mechanical energy of the voice coil and cone. Secondly this mechanical energy is converted into acoustic energy by the interaction of the cone with the neighbouring air molecules. Both these conversions must be accurate if the final result is to be a low distortion replica of the input voltage waveform.

The laws that apply to mechanical and acoustic forces are directly analogous to those of electrical forces and for this reason we can represent what happens in any acoustic or mechanical problem by a circuit diagram. In mechanics and acoustics there are direct and simple relationships like Ohm's law in electronics. It is only the complex arrangement of mechanical or acoustic circuit elements that makes the picture look complicated. ▶

FIGURE 3A

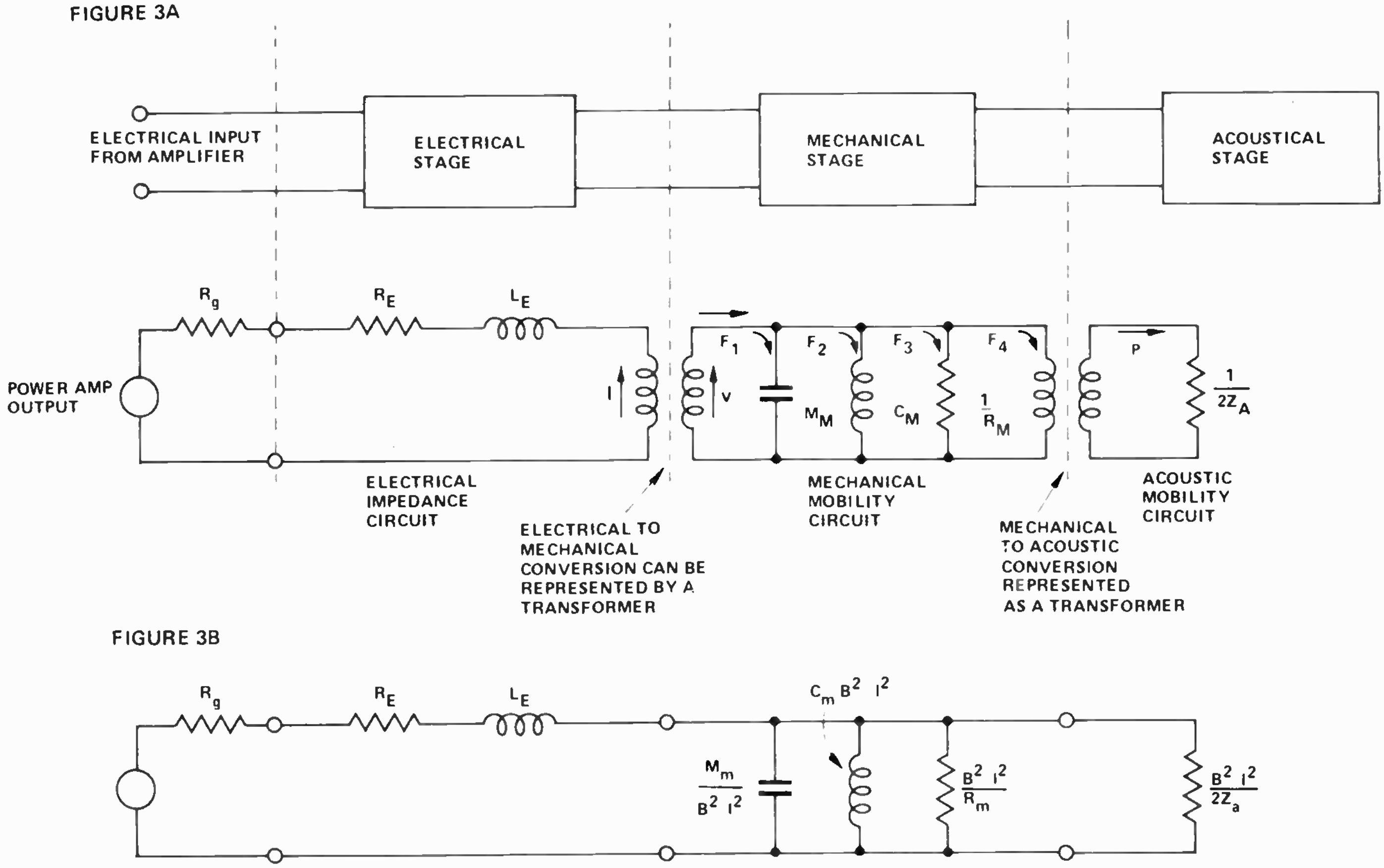


Figure 3. **Equivalent circuit for a typical moving-coil direct-radiating driver mounted in an infinite baffle.**

Just as an electronic circuit can look complex, but can be broken down into smaller and simpler circuits, so too can any acoustic or mechanical problem.

We can represent a complete picture of a dynamic loudspeaker by a circuit diagram showing electrical/mechanical and mechanical/acoustic conversions (see Figure 3a).

The power amplifier is connected via a net series resistance Rg, to the terminals of the loudspeakers. This resistance is the result of the internal resistance of the power amplifier and connecting cables. Since the voice coil is a coil of wire it possesses both inductance and resistance. The applied electrical signal sees these two in series and we represent this by the resistance R_E and the inductance L_E. The "E" simply implies that these are electrical quantities. Current flowing in the voice coil gives rise to the magnetic field that causes mechanical movement of the voice coil and cone assembly. This conversion of electrical to mechanical energy is represented in the circuit diagram as a transformer. Voltage across the primary is represented by the letter "e" and gives rise to velocity 'v', of the voice coil and cone assembly at the secondary of the transformer.

The total force applied by the voice coil ("F") is shown in the mechanical stage as "flowing" through the "wires" just as current would flow through the wires of an electrical circuit. This force sees three mechanical components in parallel, a mechanical capacitance M_M, a mechanical inductance C_M and a mechanical resistance, $1/R_M$. The mechanical capacitance M_M is caused by the mass of the cone. As frequency rises inertia comes into play and it becomes increasingly difficult for the cone to follow the input voltage waveform. The mass of the cone causes a frequency response roll-off at higher frequencies. This could be represented either by an inductance in series or a capacitance in parallel with the load. In Figure 3 this has been shown as the parallel capacitance, M_M.

A loudspeaker cone has a certain springiness, due to the nature of the cone's suspension and the overall construction of the particular driver. We specifiy this springiness by a spring constant, which is simply a number, represented by the letter 'k'. In loudspeaker technology we more often use the term compliance rather than spring constant. Compliance C_M, is defined as:

$$C_M = \frac{1}{k}$$ where 'k' is the spring constant.

The compliance impedes large movement of the cone. Since bass frequencies require longer cone excursions the compliance of the driver causes a frequency response that falls as frequency decreases. This can be represented as a capacitance in series or an inductance in parallel with the load. In Figure 3 the compliance C_M is represented as an inductor in parallel with the load.

The remaining term in the mechanical part of the loudspeaker circuit diagram is the mechanical resistance. Just as all circuit elements in an electronic circuit have resistance, so to does the mechanical circuit. The resistance is seen in series with the whole mechanical circuit and could be represented as a series resistor or a parallel inverse resistance. If R_M is the mechanical resistance of the circuit, an inverse resistance is defined as:

$$\frac{1}{R_M}$$

In Figure 3a force is shown as 'flowing' in the mechanical 'wires'. The total available force is shared into four major parts; the forces needed for the mass M_M, the compliance C_M, the mechanical resistance R_M and the load. If we define these four forces as F_1, F_2, F_3 and F_4 respectively, we can say that

F (the total force available) = $F_1+F_2+F_3+F_4$, and this has been shown in the mechanical circuit diagram. We have to represent the series resistance as an inverse resistance, $\frac{1}{R_M}$ and place it in parallel with the load, to illustrate the way it obtains its part of the total available force (F).

In this case the load is the primary of the mechanical/acoustical transformer. Of course this transformer doesn't actually exist. It is merely a way of representing the conversion of mechanical energy to acoustic energy by the interaction of air molecules with the surface of the loudspeaker cone. Mechanical force in the primary of the transformer is converted into sound pressure 'p', in the acoustic circuit.

In Figure 3a it is assumed that the loudspeaker is mounted in an infinite baffle. This is a partition that extends to infinity in all directions, cutting the universe into two halves, with a hole in which the loudspeaker is mounted. This is just a little impractical, but the only important thing is that no sound produced by the back of the speaker cone can interact with the sound from the front.

In order to move air molecules, the cone must do work so the air impedes movement of the cone. This impedance is called the acoustic impedance and is represented in the circuit diagram by Z_A. Since the loudspeaker is mounted in an infinite baffle the acoustic impedance is the same on both sides of the cone and becomes $2Z_A$.

We are now in a position to understand the causes of variations in the electrical impedance and acoustic radiated power. As was shown earlier, the back EMF is one of the dominant forces acting to increase the driver's impedance. It is related to the velocity of the loudspeaker cone as was indicated by Equation 2. If the motion of the cone is impeded, i.e: if the cone is held, the velocity must decrease, causing a decrease in the amount of back EMF. The decreased back EMF will cause a drop in loudspeaker impedance. So an *increase* in mechanical impedance causes a *decrease* in electrical impedance. With this in mind the electrical/mechanical/acoustic circuit diagram of Figure 3a can be converted into the all electrical circuit diagram of Figure 3b.

This circuit predicts the impedance characteristics of the driver. A generally increasing impedance with frequency is caused by L_E, while M_M, C_M, and R_M form a damped parallel resonant circuit. We would expect a sharp increase in impedance at one frequency, dropping to the dc resistance of R_E and R_G and then slowly rising as frequency increases. This is exactly the response as shown in Figure 2 which is the measured impedance response of a typical 12 inch (300 mm) woofer. This resonance point is called the fundamental resonance of the driver, and being a function of the compliance of the driver, can be expected to decrease in frequency a little as the driver wears in. This is the reason some loudspeaker experimenters "run in" the driver before measuring resonant frequency.

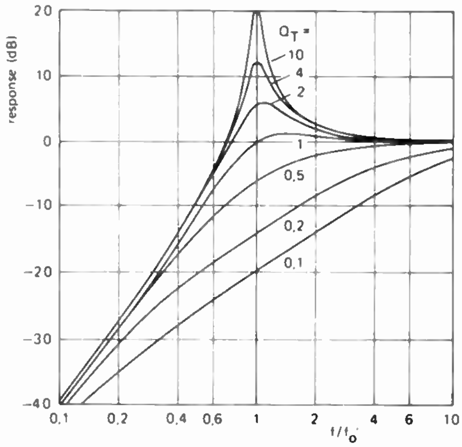


Figure 4. **Normalized frequency response of a typical low frequency loudspeaker for different values of Q_T (after Hermans & Hull, Electronic Applications Bulletin, Vol. 35, No.2., Feb. 1978, Philips).**

A more accurate model

The model of the loudspeaker developed so far has assumed that the shape of the loudspeaker cone remains unchanged and moves as a "rigid piston", following the input signal. This rigid piston theory works well at predicting the characteristics of drivers at low frequencies. At higher frequencies inertia again comes into play and the cone can no longer be considered as a rigid piston. If the driver remained a rigid piston throughout the audio spectrum its frequency response would fall off at a rate of 12 dB/octave at higher frequencies, limiting its useful frequency range.

The equation showing the relationship between the frequency of a sound and its associated wavelength is

$\lambda \nu = V_A$ where V_A is the velocity of sound in air
λ is the wavelength in meters
and ν is the frequency in Hertz.

The equation shows that the wavelength of sound decreases as frequency is increased. It should be noted that the velocity of sound depends on the medium in which the wave is propagating. The velocity of sound in the loudspeaker cone will be substantially different to that in air. Using this equation we can calculate the frequency at which the wavelength of sound approaches the radius of the loudspeaker cone. For a 300 mm (12 inch) loudspeaker this frequency is around 400 Hz and it is at this frequency that the rigid piston theory starts to come unstuck. Above this frequency the sound wave propagates up the cone, hopefully to be damped in the rubber surround. The sound wave is attenuated as it moves through the cone, and this attenuation effect increases with increasing frequency, causing a decrease in the effective cone diameter. This is the effect that enables a single cone loudspeaker to operate over a wide frequency range, since the decreasing effective cone diameter decreases the inertia presented to the coil assembly at higher frequencies. It should be noted that, in this range of the frequency spectrum, the rim and the cone will be radiating in anti-phase with the coil assembly. The way the cone material and suspension react to this multiple wave propagation is one of the biggest differences between a good driver and a poor one. It is for this reason that metal cones for instance are so often unsuccessful. Their ability to damp multiple resonances is generally poor in comparison to materials like paper or plastics.

Damping and Q-factor

In midranges and tweeters the drivers can be operated in frequency ranges that exclude their fundamental resonances. The crossover points are usually chosen so that at least one full octave exists between the crossover point and the fundamental resonance. In the case of bass drivers however it is necessary to operate the driver at and below the resonance of the woofer.

This is the main reason so many different bass loading principles have been developed. The fundamental resonance of the bass driver must be damped so that an acceptably flat frequency response can be established. If the resonance is not damped adequately, the all too common 'one note bass' sound results. This is a particularly noticeable and fatiguing loudspeaker fault and considerable effort must be spent on obtaining a smooth bass end response.

Since the loudspeaker is a resonant circuit the amount of damping can be specified by quoting the Q or quality factor. Q is defined by:

$$Q = \frac{f_o}{f_1 - f_2}$$ where f_o is the frequency of the fundamental resonance. and f_1, f_2 are the 3 dB points.

Figure 4 shows a graph of bass-end frequency responses at a variety of Qs. Although the flattest response appears to be given by the case when the Q=1, this is not the optimally damped case and some boomy bass often occurs in bass systems with Qs around unity. The best Q is probably about 0.5. The bass is not boomy but is also not over-restricted which can happen if the Q drops to around 0.2 or 0.3. The best damping for any specific case needs to be established by experiment and ultimately, as always, the ear must be the final test.

Loudspeaker compliance, the total mass of the cone and the net series resistance with the voice coil, all determine the response of any loudspeaker system and any or all of these can be adjusted in order to achieve the optimum damping and frequency response. In practice, adjustments to the Q of the system are done by modifying the compliance and acoustic mass and resistances caused by the loudspeaker enclosure rather than modification of the driver itself.

The enclosure

The circuit in Figure 3 has been developed assuming that the driver is mounted in an infinite baffle. The air load on the cone of the loudspeaker is represented by an impedance of value:

$$\frac{1}{2Z_A}$$

When the driver is mounted in a practical loudspeaker enclosure, this acoustic impedance becomes a little more complicated and the circuit in Figure 5 replaces the simple resistor of Figure 3. This new impedance is made up of two major components. The radiation impedance from the front of the box (M_{AR}, R_{AR}) is related to the size of the baffle and is independent of the volume of the box.

The volume of air that the driver has access to is related to the effective radiating area of the cone and to the size of the baffle. Since bass frequencies have a greater dispersion than higher frequencies the volume of air accepting the radiated sound increases at lower frequencies. So the impedance on the front of the cone will be greatest at higher frequencies. If the size of the baffle is large in comparison to the radiating size of the driver the box approximates an infinite baffle down to a lower frequency than it would otherwise, and the frequency at which the driver has access to a 360° radiation pattern is decreased. This is represented by the series combination of the inductance M_{AR} and R_{AR}, which gives an impedance characteristic that increases with frequency like the front radiation.

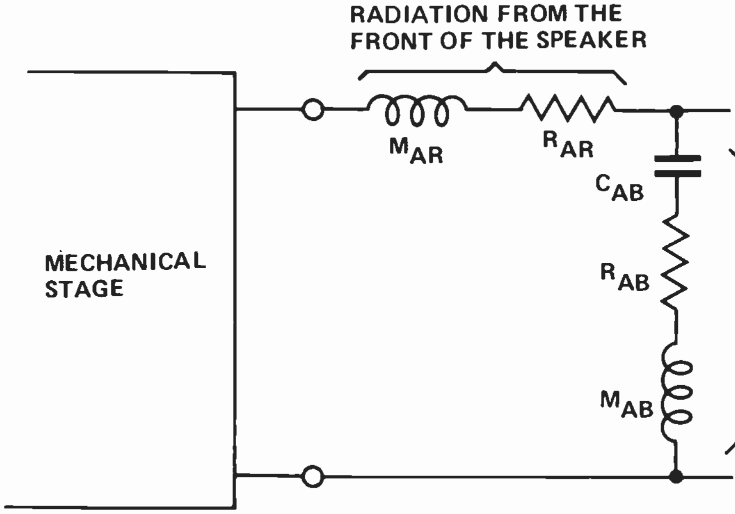


Figure 5. **Equivalent circuit diagram for the acoustic stage of a direct-radiating moving-coil driver mounted in a sealed enclosure.**

The second component of the radiation impedance is caused by the enclosed volume of air within the box. If we consider a sealed enclosure the volume of air within the box will be compressed by the driver. So the enclosure volume will affect the overall compliance of the loudspeaker system. This acoustic compliance is represented in Figure 5 by the capacitance C_{AB}. The effect of this is to increase the stiffness of the loudspeaker cone resulting in an increase in the fundamental resonance of the enclosure.

The volume of air in the box will also have an equivalent mass represented by the inductance M_{AB}. This mass will also affect the resonance of the system by increasing the overall mass of the cone. The acoustic resistance with the enclosure is shown in Figure 5 as R_{AB}.

The final resonant frequency of the driver in the box is a function of the total effect of all the compliances and masses in the system. If the total mass is represented by M_a and the total effect of the compliances is C_a then the resonant frequency of the system will be given by the equation

$$f_o = \frac{1}{2\pi\sqrt{M_a C_a}}$$

The equation shows that a decrease in either the total mass or compliance will cause an increase in the resonant frequency of the loudspeaker system.

Resonances

The acoustic circuit in Figure 5 represents the reactances caused by the enclosure around the resonant frequency, but as usual in loudspeaker science things get more complicated as frequency increases. As the wavelength of the sound wave produced inside the enclosure becomes shorter the box no longer acts as a simple spring. The produced wave travels from the back of the driver towards the rear and sides of the cabinet, where it is reflected back towards the driver. This sound wave will interact with the cone, either reinforcing or impeding the motion of the cone depending on the particular frequency. This results in successive rises and dips in the frequency response and for this reason it is important that this reflected wave is damped as much as possible. In order to absorb this unwanted energy the enclosure is usually lined with an absorptive material such as bonded mineral wool, acetate fibre or bonded hair felt.

The most important parameter of any of these materials is that they are reasonably open. If the material is too dense it will not only have little absorption but it will decrease the total available volume within the box.

Generally a layer of 25 mm speaker innerbond on the back, sides, top and bottom is about right.

Lining the box also has the effect of altering the acoustic resistance on the back of the cone. This effect is often used to increase the damping and thereby decrease the Q of the system resonance. Usually the necessary damping can be only partially achieved and it is necessary to partially fill the box with an absorptive material as well. If the enclosure is completely filled sound waves within the enclosure are converted into heat in the filler material. Normally this overdamps the box resulting in a Q sometimes as low as 0.1.

Filling the box has one other major effect. Owing to the heating of the material inside the box and its different density to that of air, the velocity of sound within the enclosure is reduced dramatically. The speed of sound in air is around 344 m/s and this could drop to as low as 292 m/s. This has the effect of increasing the compliance of the enclosure and thereby decreasing resonant frequency in the same way as increasing the box volume would. An optimally filled box could appear some 30% to 40% bigger than it really is.

Throughout this article we have discussed the sealed enclosure, leaving explanations of other bass loading techniques to a later time. One of the most common enclosures is the bass reflex, which uses a port cut in the baffle to augment the bass radiation of the driver. The acoustic circuit diagram must show the effects of the mass, compliance and resistance of the port in addition to those shown for the rest of the box, making the loudspeaker equivalent circuit even more complicated. Both the bass reflex and the sealed enclosure are capable of very good results and it is not possible to state simply which is better. We have omitted a detailed discussion of the principles of the bass reflex loudspeaker simply for the sake of simplicity. ●

References

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Vibration and sound radiation of loudspeaker diaphrams; A.J.M. Kaizer.

Loudspeakers in vented boxes: A.N. Thiele.

Principles and problems in loudspeaker design

The second article on this subject concludes with a discussion on cross-overs.

THE LAST ARTICLE dealt with the characteristics of a typical moving-coil direct-radiating loudspeaker and the interactions that are likely to occur with the loudspeaker enclosure. Once these problems are understood and the bass performance has been optimised, we are in a position to finish the design. I have discussed the bass end of the audio spectrum first not because it is the most important, but simply because it is the most difficult to optimise. The mid-range is by far the most critical part of the audio spectrum since it is midrange distortion that the ear objects to more than any other.

It was shown before that drivers have limited frequency responses and that it is therefore necessary to use several drivers, each covering its own frequency range. By far the most common arrangement is the three-way, so called because it uses three drivers to cover the audio range. A woofer covers the bass end, crossing over to a midrange driver somewhere between 400 Hz and 1 kHz. The midrange driver, sometimes called a 'squawker', carries the frequency range from this crossover point up to where the tweeter takes over, usually around 3 kHz to 5 kHz. The tweeter covers the remainder of the audio spectrum up to around 18 kHz, about the limit of human hearing. A crossover is used to separate the input signal from the output of the power amplifier into the three frequency bands.

Passive loudspeaker crossovers

The design of the crossover for any particular group of drivers must be done only after a thorough investigation into the characteristics of the drivers has been carried out. It is essential to choose drivers with an adequate overlap in their frequency responses, or a 'hole' will result in the response of the final loudspeaker. The amount of overlap needed depends on the slope of the filters used in the crossover. If a fast slope is used a smaller overlap is required, but filters with very fast slopes are complicated and expensive.

The basic crossover filter consists of a low pass and a high pass section. In a two way loudspeaker only one of these sections would be used, while in a three way loudspeaker two sections are used, one for the bass-mid crossover and the other for the mid-treble crossover. The simplest crossover is called a first-order crossover and has a slope in its attenuating region of 6 dB/octave. An octave is a range of frequency such that the highest frequency in the band is double the lowest frequency; for instance, an octave above 50 Hz is the frequency range 50 Hz to 100 Hz, while an octave above 5 kHz is 5 kHz to 10 kHz. This is not a precise definition of an octave but is essentially correct and is adequate for loudspeaker analysis.

Figures 2 and 3 show circuit diagrams for series and parallel first-order crossovers. The series configuration is less commonly used since it is only applicable to two-way loudspeakers and has no advantages over the parallel type. If the first-order crossover is terminated with ideal resistive loads, the two slopes will add to give a linear response with the phase response perfectly preserved. In this respect it is fairly unusual since no other simple passive crossover will give a response that is linear in both frequency and phase. Unfortunately 6 dB/octave slopes require a choice of drivers with very broad overlapping frequency responses. Generally it is necessary to have a usable response from a driver to a frequency where the crossover is giving about 12 dB of attenuation. For a 6 dB/octave low pass crossover this would be two octaves above the crossover point. A woofer crossing out at 500 Hz would be very unlikely to have a response to 2 kHz, so a 6 dB/octave filter could not be used.

The most common crossover is the ▶

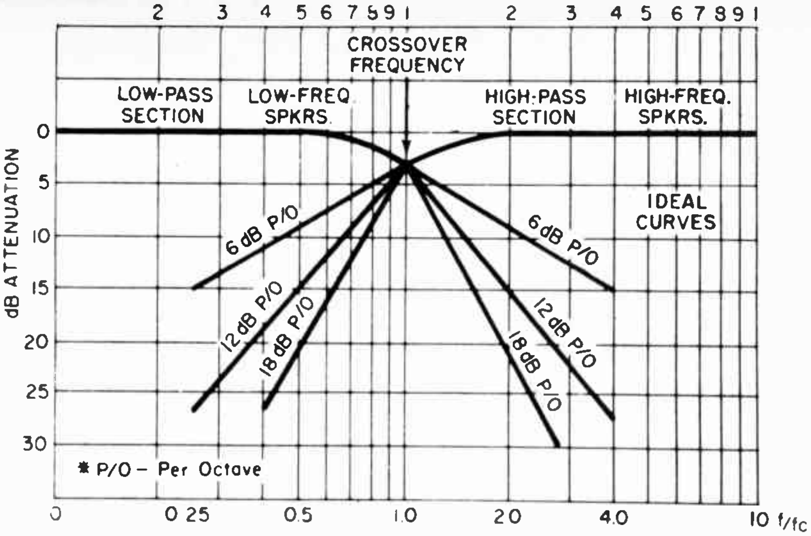


Figure 1. **Typical frequency response curves of a two-way crossover network. The graph shows three different pairs of filters, each having a different rate of attenuation.**

CONSTANT K CROSSOVER NETWORKS

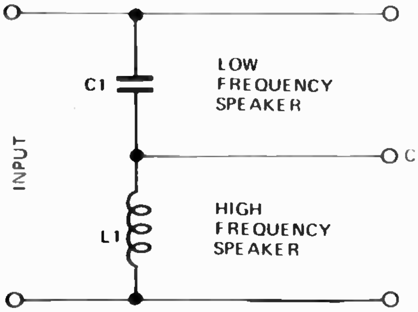


Figure 2 – **series type, 6 dB/octave.**

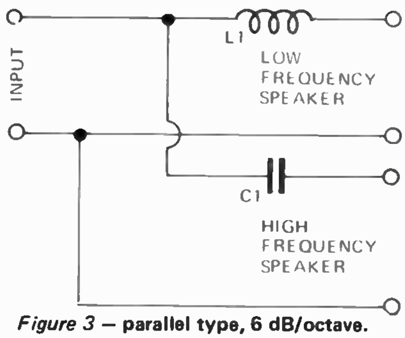


Figure 3 – **parallel type, 6 dB/octave.**

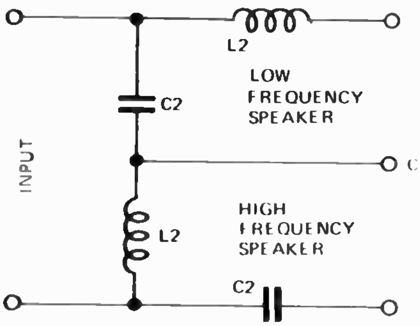


Figure 4 – **series type, 12 dB/octave.**

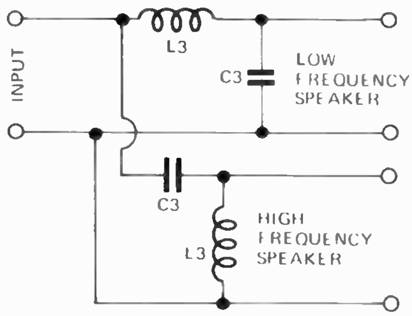


Figure 5 – **parallel type, 12 dB/octave.**

FREQ.	C1	C2	C3	L1	L2	L3
100	199	281	141	12.7	9	18
150	133	188	93.8	8.49	6	12
200	99.5	141	70.3	6.37	4.5	9
250	79.6	113	56.3	5.09	3.6	7.2
300	66.3	93.8	46.9	4.24	3	6
350	56.8	80.4	40.2	3.64	2.57	5.14
400	49.7	70.3	35.2	3.18	2.25	4.5
500	39.8	56.3	28.1	2.55	1.8	3.6
600	33.2	46.9	23.4	2.12	1.5	3
750	26.5	37.5	18.8	1.7	1.2	2.4
1000	19.9	28.1	14.1	1.27	.9	1.8
1250	15.9	22.5	11.3	1.02	.72	1.44
1500	13.3	18.8	9.38	.849	.6	1.[illegible]
2000	9.95	14.1	7.03	.637	.45	.9
2500	7.96	11.3	5.63	.509	.36	.72
3000	6.63	9.38	4.69	.424	.3	.6
3500	5.68	8.04	4.02	.364	.257	.514
4000	4.97	7.03	3.52	.318	.22[illegible]	.45
5000	3.98	5.63	2.81	.255	.18	.36
6000	3.32	4.69	2.34	.212	.1[illegible]	.3
7500	2.65	3.75	1.88	.17	.12	.24
10000	1.99	2.81	1.41	.127	.09	.18

Table 1: **Component values for constant K loudspeaker crossover networks. Inductance in millihenries, capacitance in microfarads, speaker impedance = 8 ohms.**

second order crossover, having a slope of 12 dB/octave. Figures 4 and 5 show the circuit diagrams for series and parallel second-order crossovers. Once again the series configuration is less commonly used. When terminated with ideal resistive loads the second-order crossover does not give an overall flat response. The phase characteristic causes the outputs of each half of the crossover to approach a 180 degree phase difference at the crossover point. The two outputs cancel each other, leaving a massive hole in the frequency response of the system. The 'cure' is to invert one of the drivers so that it is driven out of phase normally. The phase inversion around the crossover point brings the two drivers in phase again and the two outputs add, instead of cancelling. Unfortunately they still don't add perfectly and the result is an overall response that has a slight hump in the frequency response of around 2 dB. This is not really noticable, as few drivers have responses that are flat to this degree.

At the present time there is a great deal of discussion as to whether this non-idealistic phase response is audible. Some manufacturers insist that it is audible and design their loudspeakers accordingly, while others are most emphatic that it is not audible. The first work that I know of that was done on the subject was by Helmholtz, in his "Sensations of Tone". The quality of any sound was said to be a result only of the relative intensities of the component sine waves and not their phase relationships. The waveshape could therefore be totally different but they would sound the same.

There is another source of phase error caused by the misalignment of the acoustic centres of the drivers. The conventional way to mount the drivers is to simply bolt them to the front panel. This lines up the chassis of the drivers, but since different drivers have different depths, the voice coils of the drivers are all at different distance from the listener. If two notes are sent simultaneously to both the woofer and the tweeter for example, the note sent to the tweeter will get to the listener momentarily before the note from the woofer. Furthermore, the woofer cone is heavier than the tweeter or midrange cones and this combined with the effect of the air load on the drivers moves their actual acoustic centres even further away from the chassis. Manufacturers concerned with this effect mount the drivers on a multi-level front panel so that the tweeter is further away from the listener than the midrange. Similarly, the midrange is mounted on a plane that is further away from the listener than the woofer. This gives the sound from the midrange and woofers a head start over the tweeter, and attempts to correct for the differences in their acoustic centres.

Both types of phase errors need to be recognised and dealt with independently if a meaningful analysis of the audibility of phase errors in loudspeakers is to be carried out. Even if phase errors of this magnitude are audible (and only experiment can tell us), an extremely good loudspeaker can still be constructed along the more conventional techniques using second-order crossovers with drivers mounted on a plane baffle.

The 4000 Series of loudspeaker projects described in ETI 1980 use the more conventional approach to driver mounting and crossover design to simplify construction and decrease cost. If you choose to experiment with the audibility of phase in loudspeakers and construct a Series 4000 loudspeaker with a stepped front panel, the best way to establish the correct distance between the panels is by experiment. The drivers should be connected to the crossover and mounted in separate enclosures. The size of these enclosures is not critical.

Supply the power amp driving the loudspeaker with a source of low repetition pulses (or a low frequency square wave around 20 Hz). If the loudspeaker is now monitored with a microphone and the output of the mic amplifier fed to an oscilloscope, the transient performance of the loudspeaker can be determined.

When the front baffles of the enclosures are aligned, as would be the case in a conventional loudspeaker, the input pulses will be seen to be converted into a series of pulses. Each pulse corresponds to one of the drivers. If the enclosures are moved slowly back with respect to the woofer enclosure these pulses will merge into a single pulse. This is the correct position for the baffles, and using these measurements the final enclosure can be built. If you have the necessary equipment to do this experiment we would be interested in hearing about your results.

The crossovers described so far belong to a class of filters called constant-K filters. These filters are designed on the assumption that the product of the impedances of the capacitor and the inductor in the stage is equal to the square of its characteristic resistance, i.e.

$$Z_C \times Z_L = R_o^2$$

The characteristic resistance of a filter is that resistance into which there is maximum power transfer. Originally, 'K' was used instead of the now more common symbol R_o for the characteristic resistance, hence the name constant K. Table 1 gives values for constant-K filters for a variety of crossover points assuming an 8 ohm resistive load.

M-derived filter sections

The assumptions made to simplify the design of the constant-K filters lead to some non-ideal characteristics. It is sometimes mistakenly thought that constant-K implies constant impedance. This is not the case and the impedance is a variable with the effect occuring ▶

mostly around the crossover point. The other problem with constant-K filters is that the slopes are slowest near the crossover point. The solution to these problems has been known since 1923, when Zobel proposed that other sections could be used to flatten the response within the passband and sharpen the roll-off point. These stages are called M-derived sections, since the values of inductance and capacitance used in the filter are obtained by first deriving them for a constant-K type filter and then converting these values into M-derived values with the use of a mathematical equation that contains the term M. M is simply a number between 0 and 1, usually around 0.6 for crossover applications. Either the phase or frequency characteristics may be optimised but not both at once; 0.6 is a good compromise. Table 2 and Figures 6 to 9 give values for inductors and capacitors for M-derived crossovers with M = 0.6.

The other major advantage of this filter is that it allows a third-order or 18 dB/octave filter to be built. Third-order filters can be made to have a linear frequency characterisitic when the outputs of the two channels are summed, but like the second-order filters described before, suffer from a very non-linear phase response. Each filter shifts the phase at the crossover point by 180 degrees, so there is a 360 degree phase shift between the two outputs.

Loudspeaker impedance

So far I have assumed that the loudspeakers connected to the crossover are fixed 8 ohm loads, but as was seen in last month's article, this is most definitely not the case. Most drivers have an impedance characteristic that presents maximum impedance at their resonant frequency, dropping to the nominal dc resistance of the driver at a frequency above this, followed by a generally increasing impedance as frequency rises (see Figure 2 in the last article). Provided the driver is not being used near its resonant frequency, which should always be the case with midranges and tweeters, this impedance variation can be corrected by a series capacitor-resistor network placed in parallel with the loudspeaker Figure 10 shows a typical circuit. This network has an impedance that decreases with increasing frequency, tending to cancel the increasing impedance of the driver. The component values shown are applicable to an average woofer, although the actual values in any specific application are best established by experiment. This works very well and it is not difficult to obtain an impedance response that is flat within one ohm over most of the driver's operating range.

Matching sensitivities

Once the crossover points have been established from an analysis of the driver's best operating regions, the final step is to equalise the various sensitivites of the different drivers. This is done by a resistor divider network as shown in

M-DERIVED CROSSOVER NETWORKS

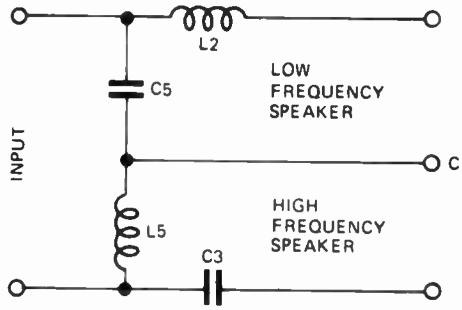


Figure 6 — series type, 12 dB/octave.

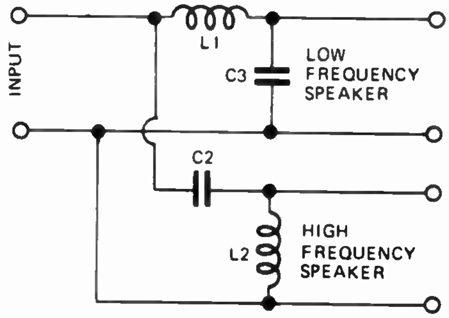


Figure 7 – parallel type, 12 dB/octave.

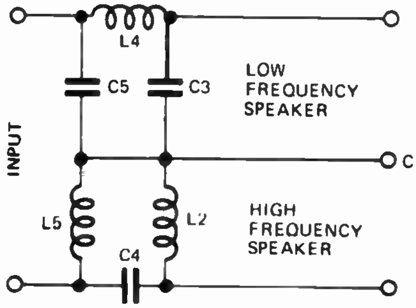


Figure 8 — series type, 18 dB/octave.

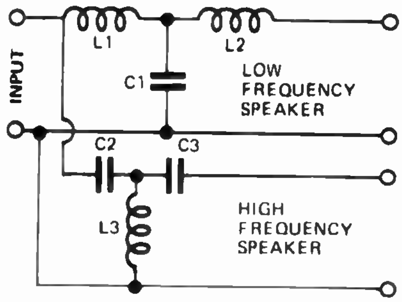


Figure 9 – parallel type, 18 dB/octave.

FREQ.	C1	C2	C3	C4	C5	L1	L2	L3	L4	L5
100	398	124	199	99·4	318	20·3	12·7	6·37	25·4	7·96
150	265	82·9	133	66·3	212	13·5	8·49	4·24	16·9	5·31
200	199	62·1	99·4	49·7	159	10·1	6·37	3·18	12·7	3·98
250	159	49·7	79·5	39·7	127	8·15	5·09	2·55	10·1	3·18
300	133	41·4	66·3	33·1	106	6·79	4·24	2·12	8·49	2·65
350	114	35·5	56·8	28·4	90·9	5·82	3·64	1·82	7·28	2·27
400	99·4	31	49·7	24·8	79·5	5·09	3·18	1·59	6·37	1·99
500	79·5	24·8	39·7	19·9	63·6	4·07	2·55	1·27	5·09	1·59
600	66·3	20·7	33·1	16·5	53	3·4	2·12	1·06	4·24	1·33
750	53	16·5	26·5	13·2	42·4	2·72	1·7	·849	3·4	1·06
1000	39·7	12·4	19·9	9·95	31·8	2·04	1·27	·637	2·55	·796
1250	31·8	9·95	15·9	7·96	25·4	1·63	1·02	·509	2·04	·637
1500	26·5	8·29	13·2	6·63	21·2	1·36	·849	·424	1·7	·531
2000	19·9	6·22	9·95	4·97	15·9	1·02	·637	·318	1·27	·398
2500	15·9	4·97	7·96	3·98	12·7	·815	·509	·255	1·02	·318
3000	13·2	4·14	6·63	3·32	10·6	·679	·424	·212	·849	·265
3500	11·3	3·55	5·68	2·84	9·09	·582	·364	·182	·728	·227
4000	9·95	3·11	4·97	2·49	7·96	·509	·318	·159	·637	·199
5000	7·96	2·49	3·98	1·99	6·37	·407	·255	·127	·509	·159
6000	6·63	2·07	3·32	1·66	5·31	·34	·212	·106	·424	·133
7500	5·31	1·66	2·65	1·33	4·24	·272	·17	·085	·34	·106
10000	3·98	1·24	1·99	·995	3·18	·204	·127	·064	·255	·08

Table 2: **Component values for M-derived loudspeaker crossover networks. Inductance in millihenries, capacitance in microfarads, M = 0.6, speaker impedance = 8 ohms.**

A Bruel and Kjaer sound level meter. Instruments like this are used to determine the frequency response of a loudspeaker. This instrument has several weighting curves built in as well as a one octave filter set to allow pink noise analysis.

Figure 11. A simple resistor placed in series with the driver would of course decrease the power in the loudspeaker for a given signal voltage, but this increases the impedance seen by the rest of the crossover altering the crossover frequency point. The resistive divider network shown in Figure 11 can be set to represent a fairly constant 8 ohm load. Resistor R2 is placed in parallel with the driver, resulting in a decreased total impedance. This impedance is then brought back up to the desired impedance by placing R1 in series. The correct values for R1 and R2, assuming an 8 ohm loudspeaker system, are given by the following three simple equations.

1. $d = \text{antilog} \dfrac{-\text{signal drop in dB}}{20}$

2. $R2 = \dfrac{8d}{1-d}$

3. $R1 = \dfrac{64}{8 + R2}$

First establish the amount of attenuation that is required in dB. Normally by this stage a frequency response curve has been established by measuring the loudspeaker, and an estimate of the required attenuation can be obtained from this. Now use equation 1 above. The antilog of a number can be found either using log/antilog tables or the inverse log key on any scientific calculator. I have used the symbol 'd' for the result of equation 1 mainly to simplify the written form of equation 2, but in reality 'd' is equal to the voltage across the loudspeaker divided by the voltage from the amplifier. i.e.

$$d = \frac{V_s}{V_i}$$

where 'V ' is the signal voltage across the loudspeaker and 'V_i' is the applied signal voltage from the amplifier.

The result of equation 1 is plugged into equation 2, which yields the correct value for R2. The value of R2 is then used in equation 3 to obtain the value of R1. For example, if a midrange is to be decreased in sensitivity by 3 dB, equation 1 becomes:

$$d = \text{antilog} \frac{-3}{20}$$

$$d = \text{antilog} -0.15 = 0.7079$$

So a 3 dB drop in output signal level is equivalent to decreasing the signal voltage across the loudspeaker to $0.7079V_i$. Plugging this into equation 2 gives

$$R2 = \frac{8 \times 0.7079}{1 - 0.7079} = 19.4 \text{ ohms.}$$

Using this result in equation 2 gives

$$R1 = \frac{64}{8 + 19.4} = 2.3 \text{ ohms.}$$

The nearest value resistors to these would be 18R and 2.2R, and with these resistors the impedance presented to the remainder of the crossover would be approximately 7.7 ohms which is close enough in practice.

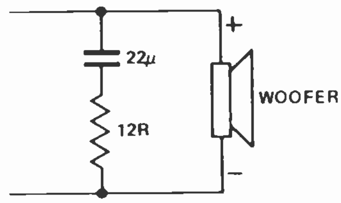


Figure 10. **Circuit to improve the apparent impedance of a loudspeaker.**

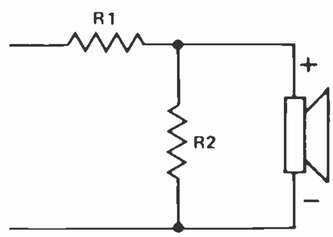


Figure 11. **Potential divider used to compensate for different loudspeaker sensitivities.**

The final test

Now that the bass performance has been optimised through a suitable choice of enclosure size and damping, the crossover points and slopes have been established, and the impedance and sensitivities of the various drivers corrected, the final subjective tests can begin. The importance of good subjective testing in loudspeaker design cannot be overestimated. The most common form of subjective analysis, other than simply listening to some good records, is an A-B test with other loudspeakers. Although this method can give some meaningful results, its validity is generally overestimated in my opinion. The best form of subjective testing is comparison to the original live performance. Simply recording a voice onto high quality recording equipment with a good microphone will tell you more about a loudspeaker than any amount of A-B testing. ●

The ins and outs of solar cells

Energy derived from fossil fuels — the world's major source of energy today — was originally provided by the sun, converted by photosynthesis with an efficiency of about 0.025%! Compared to modern solar cells, which have an efficiency of about 12%, we're on a real loser with fossil fuels. At the moment, they're convenient — but they won't always be so.
Here is a short practical guide to solar cells, their uses and abuses.

WE HAVE ALL BECOME vitally concerned about our energy resources, and rightfully so. Most people see the energy crisis in terms of paying more for a tank of petrol, but the implications run much deeper than that. Just think how many commodities are based on the oil industry – the pen I use to write with is plastic, the table top is plastic veneer, even the carpet is synthetic – all made from petroleum products.

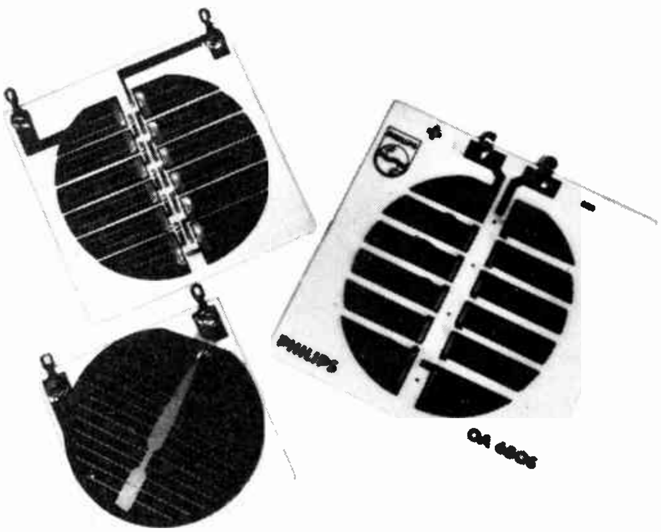
A very large percentage of our business trade is in oil-based products e.g: clothing, photography, medicine, and household goods, to mention just a few. In fact Western economies are based so heavily on oil products that, if anything suddenly happened to the supply, most western nations would collapse.

An enormous amount of energy is radiated by the sun. It is, in fact, our primary energy source. On a clear day the Earth receives about one kilowatt of solar energy per square metre on its surface. About 30% is reflected back into space, 47% is converted into heat, the rain cycle uses another 23% (which can be tapped to provide hydro-electric power in suitable mountainous areas) while wind, waves, and convection currents account for about 0.25%.

The remainder, about 0.025% (!), is stored by photosynthesis in plants. It is this energy that eventually goes to make coal, oil, and shale oil. The energy derived from petroleum which we use so extensively today is the accumulation of this trickle of energy into photosynthesis over millions of years. No wonder it's running out!

In fact it has been estimated it would take six million years of photosynthesis to provide us with an extra six months of oil and coal!

Solar energy can be harnessed in many different ways. Hydro-electric power is a result of the rain cycle; thermal gradients in tropical oceans have been used in an experimental generating station off Cuba as long ago as 1929; wind power is showing promise with experimental generating stations using large windmills and solar collectors have been devised to capture some of the heat which would otherwise be re-radiated and lost, converting it to hot water for domestic and commercial heating.



A set of experimental 'demonstration' solar cells made here by Philips at their Hendon, S.A., plant.

What solar cells offer

Solar cells offer a much brighter future (. . . pardon the pun) as a source of electrical energy. Firstly, they provide energy in a clean, transportable, convenient form – electricity. The predominant source of energy for electrical generation today comes from fossil fuels and hydro-electric schemes. A very few generating schemes use hydrothermal energy from natural hot springs.

Secondly, solar cells can provide energy very close to the point of consumption without requiring the transmission of energy across a distance or replenishment of fuel. Very handy in isolated locations.

Thirdly, they're relatively efficient . . . and they have a long life.

One shouldn't forget, too, that they are made from the most common substance on Earth – silicon.

To date, the most extensive use of solar cells has been in space. They have been employed as power sources for satellites for many years. Research has improved the efficiency of solar cells over the years, and the position is likely to steadily improve with continuing research.

Solar-power satellites are currently being studied (see ETI April 1979). It is proposed to assemble huge solar cell arrays in space and beam the energy back to Earth via a high power microwave transmission, enormous antennas ('rectennas') on Earth converting the microwave energy directly to electricity for distribution.

Terrestrial use of solar cells has expanded rapidly in the last few years. Remote telecommunications installations seem to be making the greatest use of the advantages offered. Some radio amateur VHF repeater stations in

TYPICAL VOLTAGE-CURRENT CHARACTERISTICS C 200

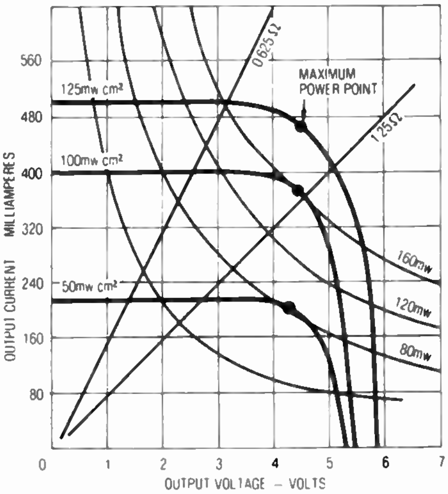


Figure 1. **Typical voltage/current characteristics of a solar cell. (Sensor Technology, type C200, distributed by Amtex Electronics).**

Australia employ solar cells to maintain charge in storage batteries used to power the installation. They are also used to charge batteries on ocean-going yachts. So you can see that hobbyists as well as professionals have been getting into the act.

Solar cell characteristics

The voltage/current characteristics of a typical single solar cell are illustrated in Figure 1. Power output contours are also shown.

At low loads (relatively high load resistance), output from the cell will be pretty nearly a constant voltage – around 0.55 V to 0.6 V – depending on the amount of energy received. If the load is increased (by reducing the load resistance), output current (and load power) will increase in proportion until a point is reached where the output voltage rapidly 'turns over', dropping sharply if the load resistance is further decreased. In this region, the load current will remain virtually constant. Maximum power output, for a given level of energy falling on the cell, occurs at the 'knee' region of the characteristics.

The performance of a solar cell depends on the spectral distribution of the irradiation impinging on it, thus, the amount of power per unit area falling on a solar cell is not a measure of the total irradiation. The term *insolation* is used to specify both the amount of power and the spectral distribution of radiation falling on a solar cell.

The relative spectral response of a typical solar cell is illustrated in Figure 2. Part of the efficiency loss in solar cells results from the fact that their spectral response does not match the spectral output of the sun. Further energy is lost in the unused excess of energy of the absorbed photons. Conversion efficiencies at an insolation of 1 kW/m^2 (100 mW/cm^2) for typical solar cells ranges between 8% and 12%.

Solar cells arrays

The most convenient way to obtain power from solar cells is to mount a number of them in an array and connect them so as to provide a useful voltage at some convenient current or power rating. Accordingly, manufacturers make 'panels' of solar cells, constructed by encapsulating individual cells in silicon resin between two plates of glass, generally with an extruded aluminium surround for the edge, with the cells connected in series. The glass plates are chemically hardened (tempered) and made very smooth to reduce the build up of dust or other residues. This is especially important where the panels are used in remote locations.

Since most of the energy falling on the panels is converted to heat and lost, the panels have to be able to conduct the heat away by convection (primarily) or conduction. Some panels are ▶

RELATIVE SPECTRAL RESPONSE

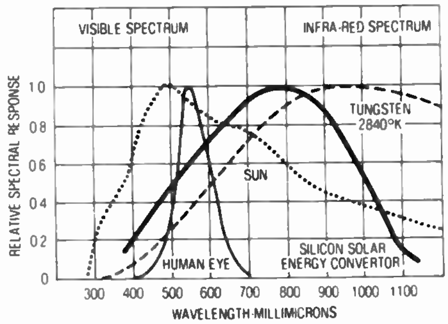


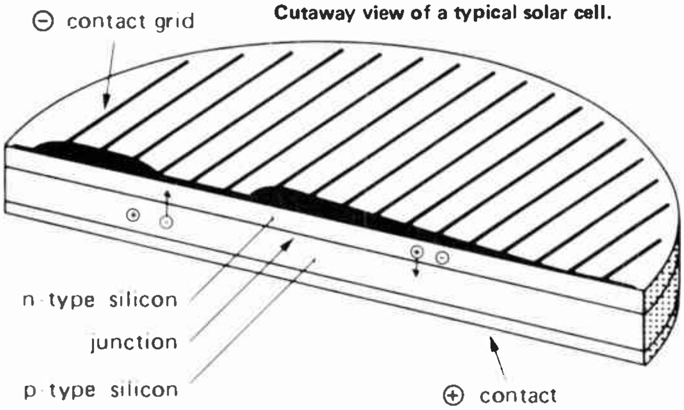
Figure 2. **Relative spectral response of a solar cell. Efficiency would be better if the response matched the Sun's output more closely.**

SILICON SOLAR CELL – HOW IT WORKS

A SOLAR CELL can be considered as a large-area silicon diode. Because it consists of a p-n junction, the junction will have a barrier potential associated with it (harking back to your diode theory) when no radiation falls on the cell. There will be an excess of electrons on the n-side of the junction (supplied by donor atoms from the doping material), some of which will diffuse across into the low electron density region on the p-side of the junction. This diffusion leaves ionised donor atoms ('holes') which create a positive space charge in the n-region close to the junction. The electrons which diffuse into the p-region will find acceptor atoms and will no longer be free to roam. This creates a negative space charge near the junction. That's how the barrier potential comes about. But, you won't be able to measure it.

The barrier potential, V_B, can be thought of as a contact potential. If contacts are made to the p-region and the n-region (with the same metal) and a high impedance voltmeter connected, no voltage will be measured. The contact potentials will cancel. Looking at the diagram, with no light falling on the cell, V_B will typically be –0.7 V, V_{C1} +0.5 V and V_{C2} +0.2 V. Hence, you won't read a thing on the meter.

Cutaway view of a typical solar cell.



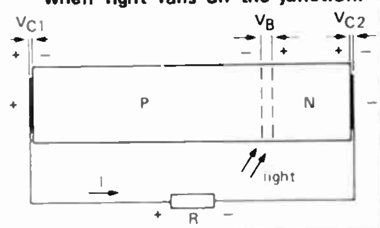
If the cell is now irradiated with light, electron-hole pairs will be generated in the junction region, separated by the field associated with V_B, the holes being forced to the p-side and the electrons to the n-side. i.e.: they move across the junction. Consequently, the barrier potential will fall considerably, to say 0.1 V!

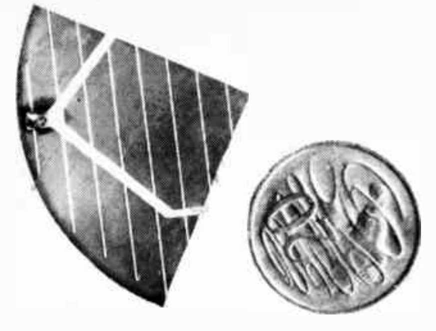
However, the p-contact will then be at a potential 0.6 V above that of the n-contact. Now, you can measure this! With sufficient irradiation, electrons charge across the junction from the p-region to the n-region – via a load and round again if you want the solar cell to do work.

Thus, convential current flow will be from the p-contact (which becomes the positive terminal) to the n-contact via a load. The maximum current obtainable is approximately proportional to the level of irradiance and the area of the cell.

Conversion efficiency of solar cells ranges between 8% and 15%, typically 10-12%, under a standard solar irradiance of 1 kW/m2 (100 mW/cm2). It is limited by three main factors: firstly, only part of the Sun's available spectrum is used: second, the absorbed photons have an unused excess of energy and lastly, some of the electron-hole pairs created are lost through recombination.

Representation of a solar cell showing the contact and barrier potentials. V_B falls considerably when light falls on the junction.





A solar cell 'piece' from Sensor Technology, type C202, used in projects in this issue.

provided with a sturdy, cast aluminium frame at the rear which serves as a heat dissipator for the array.

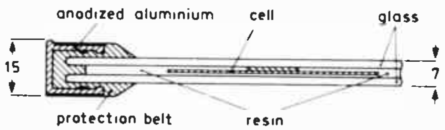
High temperatures on a solar cell panel have to be avoided, otherwise damage may result. Although individual cells can withstand quite high temperatures before they suffer structural damage, the resin potting compound cannot. Excessive heat induces strains in the resin, causing it to tear away from the surface of the cell, leaving a gap, and decomposition of the resin due to excessive temperatures can cause discolouration. The results of these two effects combine to attenuate the light falling on the cell, decreasing its efficiency.

It is important that solar panels are used within the Safe Operating Area Limits (SOAR) given in manufacturers' data. Most panels are designed so that, when used singly – for charging a storage battery, for example – they cannot be damaged. Series and parallel connection requires care to avoid excessive dissipation in particular cells. Notes on avoiding problems are given a little later in the article.

Load considerations

Operating solar cell arrays into a fixed load resistance is not ideal since, at different levels of insolation, the output voltage and current will vary and thus the maximum power output point varies. Thus, the optimum load resistance should be different for different levels of insolation. If a secondary battery (an accumulator – such as a lead-acid or nickel-cadmium type) is used as a load, this problem is largely overcome.

As an example, let's examine the characteristics of a typical solar panel – the Philips BPX47A, Figure 3. It delivers a maximum power output of almost 10 watts at a peak insolation of 1 kW/m² into a load resistance of 20 ohms. At half that insolation level (500 W/m²), power in a 20 ohm load would only be 2.9 watts. For a 12 volt accumulator (see the 'battery load line'), power delivered to the battery at peak insolation would be a little under 10 watts, but at 500 W/m² insolation it would be 4.8 watts.



Construction of the BPX47A solar panel

For this reason, solar panels are manufactured with the correct number of cells to charge a (nominal) 12 V storage battery (34 in the BPX47A). The solar cells are able to work at near-optimum efficiency and the storage batteries can provide peak demands of the power-consuming equipment and bridge overcast periods and night time when the panel receives little or no energy.

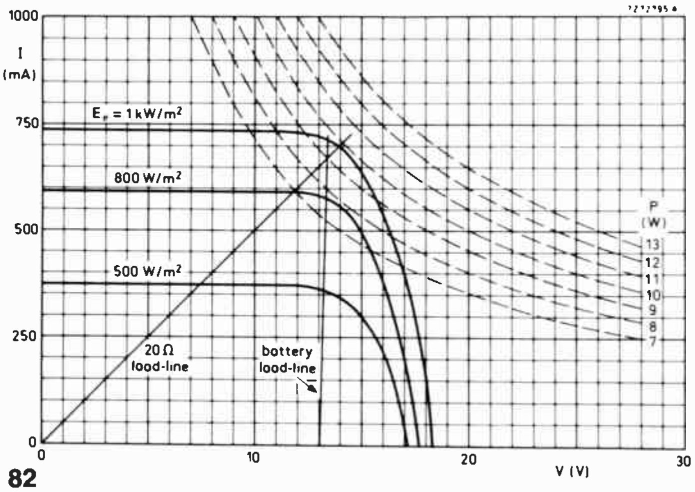
Series connection of solar cells

Any number of solar cells may be connected in series to give a desired output voltage. There are however, some points to remember. If all but one of the cells are in shadow, the irradiated cell will not be able to overcome the barrier potentials of the shadowed cells (since all their barrier potentials are in series) and no current will flow. Taking that a little further, sufficient cells in a solar array must receive irradiation so that the barrier potentials of the remaining cells can be overcome. In the extreme case, what happens when only one cell in an array does not receive sufficient irradiation? The irradiated cells will then force a current through it and the cell will develop a *reverse* voltage across it and thus dissipate power. The actual dissipation will depend on the amount of shadowing. If the irradiance to shadowed cell increases, the power dissipated will *increase* as more current will be able to flow through it, but until the cell can produce the same current as the others – by receiving the same irradiation – it will remain reverse biased.

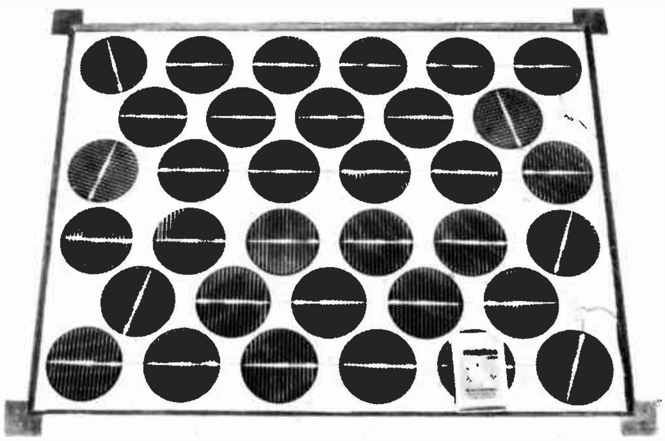
The maximum dissipation of a cell is limited by its area. As a guide, the dissipation should be less than the maximum power received at an isolation of 1 kW/m². For example, the area of one cell in the Philips BPX47A is 26 cm² and thus the maximum dissipation is 2.6 W. For the Sensor Tech. C200 (characteristics given in Figure 1), which has an area of 20 cm², maximum dissipation is 2.0 W.

An effective way of limiting the dissipation is to place a protection diode across each cell to short out any reverse voltage across the cell. This is a rather expensive solution and is unnecessary if the cells are used to charge a battery as the constant voltage characteristic of the battery will limit the maximum voltage which can be developed across any one cell. This is another reason why solar panels are designed to feed a storage battery. If however, several panels are connected in series a protection diode must be connected across each panel to limit the maximum reverse voltage.

Figure 3. Characteristics of Philips' BPX47A solar panel.



Philips' BPX47A solar panel (matchbox for size comparison.)



Parallel connection

If cells are connected in parallel to supply a higher current the voltage across each cell will obviously be the same. However if one cell receives less insolation than the others, the shadowed cell will be biased into its forward region and current will be forced through it from the other cells receiving full insolation.

In the worst case one cell in a parallel-connected array will be shadowed and the rest will receive full light. All the energy from the irradiated cells will be dissipated in the shadowed cell and it will heat up. For this reason individual cells should not be connected in parallel.

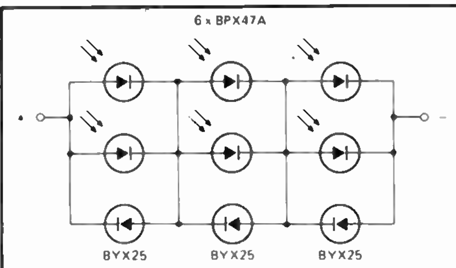


Figure 4. **Matrix connection of a solar panel to improve output. The BYX25s serve as protection diodes.**

When solar panels, or chains of series-connected cells are connected in parallel, the dissipation in a shadowed panel will be equally divided between each of its cells. In the case of the BPX47A panel with 34 cells in series the temperature rise is limited to such an extent that up to 12 panels can be safely connected in parallel.

Solar panels in series and parallel

For higher voltages and higher currents a number of solar panels can be connected in a series-parallel combination. To limit the dissipation in any panel a matrix is used as shown. With the Philips BPX47A panel, for example, the matrix must be three series by two parallel. Protection diodes are still required across each panel to limit the dissipation in individual cells; Figure 4 shows how. ●

We are indebted to Ampex Electronics and Philips for assistance with this article.

EXPERIMENTING WITH SOLAR CELLS

There are a number of interesting and instructive little experiments you can perform with solar cells. There are a number of small hobby-type electric motors around which require only 100 mA, or less, which run quite happily from 1½ V. Four Sensor Tech, C202 cell pieces or Dick Smith Z-4820 cells, connected in series, will power one of these motors. Why not convert a small battery-driven toy?

Electroplating, especially when doing it with precious metals, works best with low current density, long period operation. This method gives a beautifully smooth finish. A solar plater set-up is illustrated in the accompanying diagram.

The wirewound pot is adjusted to give 5-10 mA of current for small items, three to five times that for larger items, and the process allowed to run for three or four hours or longer, depending on the results you want. There's plenty of room for experiment here.

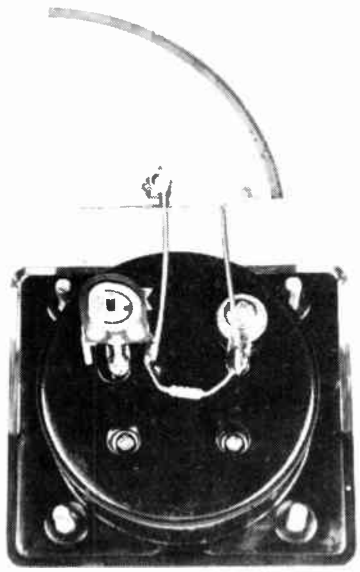
Copper plating is quite easy, and probably simplest to start out with as the ingredients are readily obtainable. The plating solution is copper sulphate and a large piece of copper wire (sanded until it's bright) will serve as the anode. Don't use a metal plating bath — remember!

Another interesting device to experiment with is a sun (or light) intensity meter. The circuit and construction details are shown here. We mounted all the bits on the terminals of a small 1 mA meter. The solar cell we used was a single Sensor Tech. C202. The device works as follows: When driving a low resistance load, the current through the load is pretty well directly proportional to the insolation (energy falling on the cell), the voltage output varying only over a small range.

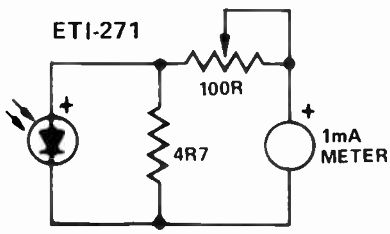
To use it, hold the device at arm's length and turn your back to the sun. Angle the unit to peak the current reading. Calibrate it by adjusting the trim pot to get a full scale reading on a bright, cloudless summer day. Full scale then represents something close to 100 mW/cm² insolation. The scale is fairly linear.

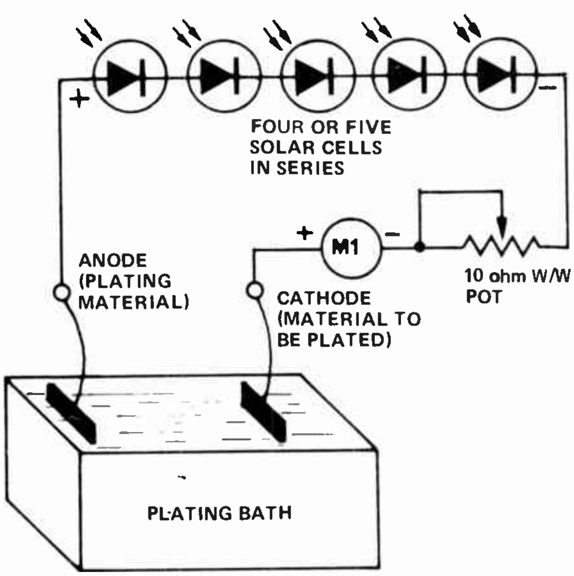
Solar cells make excellent photosensors and may be used in such applications as light-operated relays, photodensitometers, receiver for a light-beam communicator etc, etc.

Front view of the sun intensity meter we made as an experiment. The cell we used is a Sensor Tech. C202, quarter of a C200.



Rear view of the sun intensity meter showing how the cell was mounted on two pieces of 18 gauge tinned copper wire.





The Wien Bridge oscillator

The Wien bridge is probably the most popular type of low frequency sinewave oscillator. It is superior in virtually all aspects to phase shift circuits, but is not well known nor understood. This article sheds some light on this most useful circuit.

MOST STUDENTS of electronics — that includes hobbyists, you learn from your hobby don't you? — would be familiar with the "Wheatstone Bridge"; that often-handy technique for measuring unknown values of resistance. The Wien Bridge is an outgrowth of the Wheatstone Bridge. The basic circuit is shown in Figure 1.

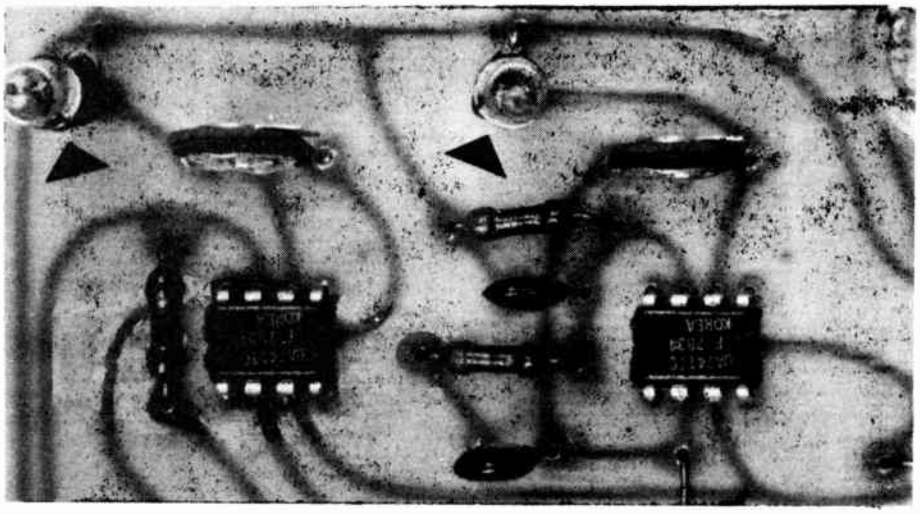
This circuit has some unique properties. The networks R1-C1 and R2-C2 form a potential divider between points A and B. Both networks have an impedance which decreases with frequency. At one frequency, and one frequency *only* (depending on the values of R1-C1 and R2-C2), the bridge will be balanced. That is, if a sinewave voltage is applied between A and B, no voltage will appear across C and D. Another interesting, and useful property of this bridge is that, at the balance frequency, the phase of the voltage across C and B will be *exactly* the same as that across A and B. The same will be true for harmonics of the balance frequency, *but*, the impedances of R1-C1 and R2-C2 will not be the same as at the balance frequency and the bridge will be unbalanced.

Well, how are these properties of the Wien Bridge used in an oscillator? The basic circuit of a Wien Bridge oscillator is shown in Figure 2. The component numbering of the Rs and Cs is the same as in Figure 1. We are assuming that the amplifier has good common-mode rejection, an infinite input impedance and zero output impedance. Fortunately, an op-amp is a reasonable approximation to this and the circuit as shown will work well with a common-or-garden 741 at frequencies up to 10 kHz.

The Wein Bridge components are connected such that positive and negative feedback is applied around the op-amp. This should be readily apparent from the way Figure 2 is drawn. The negative feedback is derived from the resistive potential divider R3 and R4. Positive feedback is provided by the potential divider R1-C1 and R2-C2. The amount of positive feedback through R1-C1 will *increase* with frequency as this network has a *decreasing* impedance as frequency increases. The parallel RC network formed by R2-C2 also has *decreasing* impedance with *increasing* frequency, tending to shunt the amount of applied positive feedback (via R1-C1) to ground. At the balance frequency, the applied positive feedback will be a maximum, falling at frequencies above and below the balance frequency. However, if the bridge is balanced, the positive feedback and the negative feedback will be equal . . . and the circuit will not oscillate. *But*, if the amount of negative feedback provided by R3-R4 is chosen to be fractionally less than the positive feedback at the balance frequency, the circuit will oscillate. Since negative feedback predominates *at all other frequencies*, and the bridge remains unbalanced, harmonics of the balance (or resonant) frequency are suppressed and the waveform produced will be a sine wave of great purity.

In practice it is necessary to include some means of sensing the amount of negative feedback so that the amplifier gain can be held at the precise amount necessary to ensure oscillation. If the amount of negative feedback is too little, the waveform will be distorted. If too much, oscillation will not occur. Secondly, if the gain varies (for whatever reason) the feedback needs to be stabilised to prevent distortion and level variations.

Twin Wien Bridge oscillator (from an ETI project) using lamps for stabilisation.



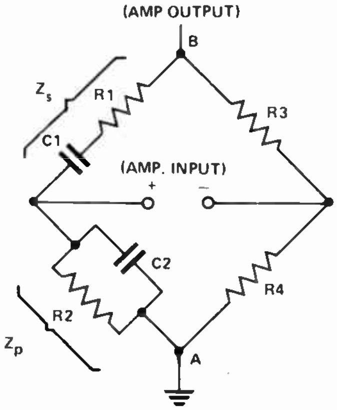


Figure 1. Basic circuit of the Wien Bridge.

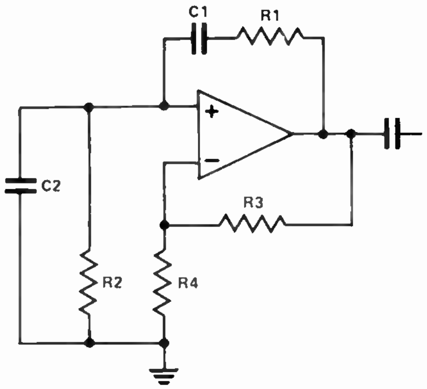


Figure 2. Basic Wien Bridge oscillator circuit.

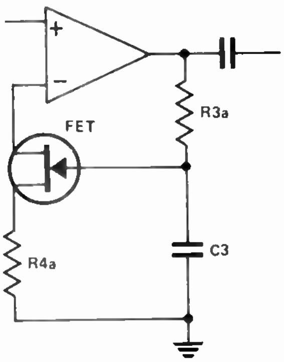
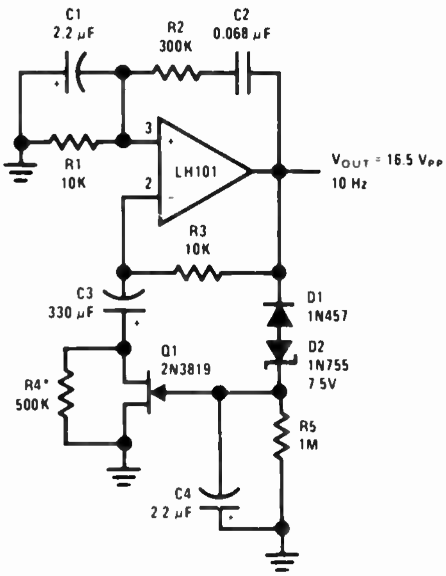


Figure 3. Feedback stabilisation using a FET.

The simplest way of doing this is to incorporate a thermistor or tungsten filament lamp in the negative feedback potential divider. If the latter is used for this purpose — and common light bulbs used for bezel lamps have tungsten filaments — it would replace R4 so that gain increases of the amplifier stage cause increased current in the lamp. This, in turn, would cause the temperature of the filament to rise, increasing its resistance, thus increasing the amount of negative feedback. The use of these temperature variable devices sets a limit on the lowest frequency at which the circuit can be used. When the period of oscillation is comparable to the thermal time constant of the particular light bulb or thermistor, the change in resistance over each cycle will bring about gain variations which result in distortion of the output waveform. Also, these devices have a "settling time" that prohibits the frequency from being changed quickly in a variable oscillator using this circuit.

Figure 4. Example of a practical Wien Bridge oscillator with a FET in the feedback (courtesy National Semiconductor).



The solution to these problems entails using a FET as part of the feedback element. The FET becomes part of R4 — as shown in Figure 3 — driven by an RC network between the op-amp output and the gate. In this way, the 'averaging time' of the circuit can be tailored to suit the job required. An example of a practical circuit is given in Figure 4.

A lot of the advantages, and the unique properties of the circuit, become apparent from a look at the mathematics involved; it's quite straightforward really.

The impedance of C1, at a certain frequency 'f', is given by:

$$Z_{C1} = \frac{1}{j\omega C1}$$

Where: Z_{C1} = impedance of C1

$$\omega = 2\pi f,$$

$$j = \sqrt{-1}$$

So the total impedance, Z_s, of the series network R1-C1 is given by:

$$Z_s = R1 + \frac{1}{j\omega C1}$$

Since the impedance of capacitor C2 is also given by:

$$Z_{C2} = \frac{1}{j\omega C2}$$

Where: Z_{C2} = impedance of C2

$$\omega = 2\pi f$$

$$j = \sqrt{-1}$$

and C2 is in parallel with R2, the total impedance of the parallel network R2-C2 (Z_p) is given by:

$$\frac{1}{Z_p} = \frac{1}{R_2} + \frac{1}{\frac{1}{j\omega C2}}$$

therefore: $$\frac{1}{Z_p} = \frac{1}{R_2} + j\omega C2$$

Oscillation will occur when:

$$\frac{R3}{R4} = \frac{Z_s}{Z_p}$$

since it is this condition which will result in unity gain.

If we let R3 = 2 x R4, and substitute this in the equations for Z_s and Z_p, this equation becomes:

$$\frac{2R4}{\frac{1}{R2} + j\omega C2} = R4\left(R1 + \frac{1}{j\omega C1}\right)$$

leading, approximately, to:

$$\omega^2 = \frac{1}{R1\,R2\,C1\,C2}$$

then $$2\pi f = \frac{1}{\sqrt{R1\,R2\,C1\,C2}}$$

and $$f = \frac{1}{2\pi\sqrt{R1\,R2\,C1\,C2}}$$

The major advantage of the Wien Bridge oscillator is its inherent stability and predictable frequency output. In other low frequency oscillators employing RC networks in the feedback, the frequency of oscillation is *directly* proportional to the values of the components in the network. In the Wien Bridge, you can see from the last equation that the frequency of oscillation is proportional to the *square root* of the component values in the network. The ease with which amplitude levelling and level stability can be achieved by using simple thermal devices in the negative feedback is another advantage. Thirdly, the low distortion possible with this circuit contributes greatly to its popularity.

On the other hand, to vary the frequency, two components have to be varied simultaneously — either C1/C2 or R1/R2. The fact that one of these is wholly 'above ground' complicates things — but it's not an insoluble problem as there are many Wien Bridge oscillators around! ●

Designing potcore inductors

Potcores offer many advantages when you need to use an inductor in a circuit. This article details a simplified design method for two common potcore sizes based on a nomograph.

FERRITE POTCORES are widely used in the construction of small inductors and transformers. However very few amateurs know how to choose a core appropriate to their needs, or how to wind a coil of specific inductance.

This article describes a simple method of designing coils for low frequency applications. The design of coils for high frequency applications, and of transformers, is beyond the scope of this article. Design details given apply only to the Philips 'P' series of pot cores and more particularly to the 18 mm (P18/11) and 26 mm (P26/16) diameter cores which are the most commonly available.

Each core size is available in four different ferrite materials (3H1, 3B7, 3D3, 4C6) to cover the frequency range from audio to about 20 MHz. Additionally each material, in each size, is available with a number of permeabilities to cover different inductance, stability and Q factor requirements.

There are two factors commonly used to classify ferrite cores. These are effective permeability (μe) and A_L factor.

The μe factor is primarily determined by the permeability of the material used and its cross sectional area, and secondly by the air gap left between the centres of the two core halves. For example an 18 mm 3B7 core without any air gap (type 0 4000) has a μe of 1750 and a tolerance on inductance of ±25%. The use of increasingly larger air gaps in the same core size and material lowers the μe but increases the stability and reduces the tolerance on inductance.

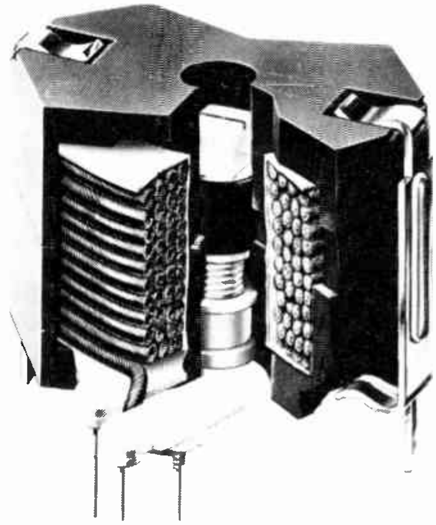
A second factor in common usage is A_L. This factor gives, in nanohenries, the inductance of ONE turn on the core. The inductance of N turns on the core is

$L = N^2 A_L \times 10^{-3}$ millihenries

The selection of a core size, a core material, a permeability value, a wire size and the number of turns depends on all the following factors:-

— inductance, stability of inductance
— frequency range
— Q factor
— unbalanced dc coil current
— level of ac coil current

Choosing the correct core taking *all* these factors into account is a difficult task indeed. However a large number of



Cutaway view of a potcore showing the winding on the internal bobbin and the inductance adjustor in the centre. The ferrite 'pot' comes in two halves, held together with a clip.

core types are eliminated by first selecting in accordance with frequency range and stability.

Frequency range

Firstly select the core material from Table 1 in accordance with the desired frequency range. To choose between 3H1 and 3B7 it is necessary to consider temperature stability.

If the tuning capacitors associated with the coil have small or varying temperature coefficients, a 3B7 core should be used as they have the lowest temperature coefficient in the range 0° - 70°C. Alternatively, if using polystrene capacitors (temp. coeff. −150 ppm/°C) a 3H1 core having an effective permeability (μe) around 150 will give excellent temperature compensation for the temperature coefficient of these capacitors.

Inductance stability

Since the inductance of a coil is

TABLE 1

FREQUENCY RANGE	CORE TYPE
0.1 — 200 kHz	3B7, 3H1
200 kHz — 2 MHz	3D3
2 MHz — 20 MHz	4C6

proportional to core permeability, the change of effective permeability (μe) with temperature determines the stability of inductance.

The percentage change of inductance with temperature is linearly proportional to μe and hence low μe cores should be used for greatest stability. Stability is therefore obtained at the expense of inductance obtainable with a given core size.

The temperature effect is not large enough to affect any but the most critical of applications and the tolerance on inductance as stated in Tables 2 and 3 will be obtained over the temperature range +15° to +35°C.

Direct current

A direct current in the winding will change the inductance value of the core and if large enough, could cause saturation. In general, large air gaps, and hence lower permeability (μe), cores should be used where large dc currents are flowing.

Q factor

The Q of a coil is influenced by different factors at different frequenceis.

At frequencies below 10kHz it is almost completely determined by the dc resistance of the winding. The Q factor of any given coil increases linearly with frequency, and the larger the core, the larger the Q. The highest Q factors are obtainable by using gapless cores of 3H1 or 3B7 material (providing that tolerance and stability are acceptable) eg. 04000 series (P18/11) and 08000 series (P26/16).

Throughout the ultrasonic range core and winding losses affect Q, but Q factors of several hundred may still be obtained by optimum choice of wire and core, such that core and winding losses are equal. For further information, on optimum design, reference should be made to Philips Data Handbook — Components and Materials, Vol. 4.

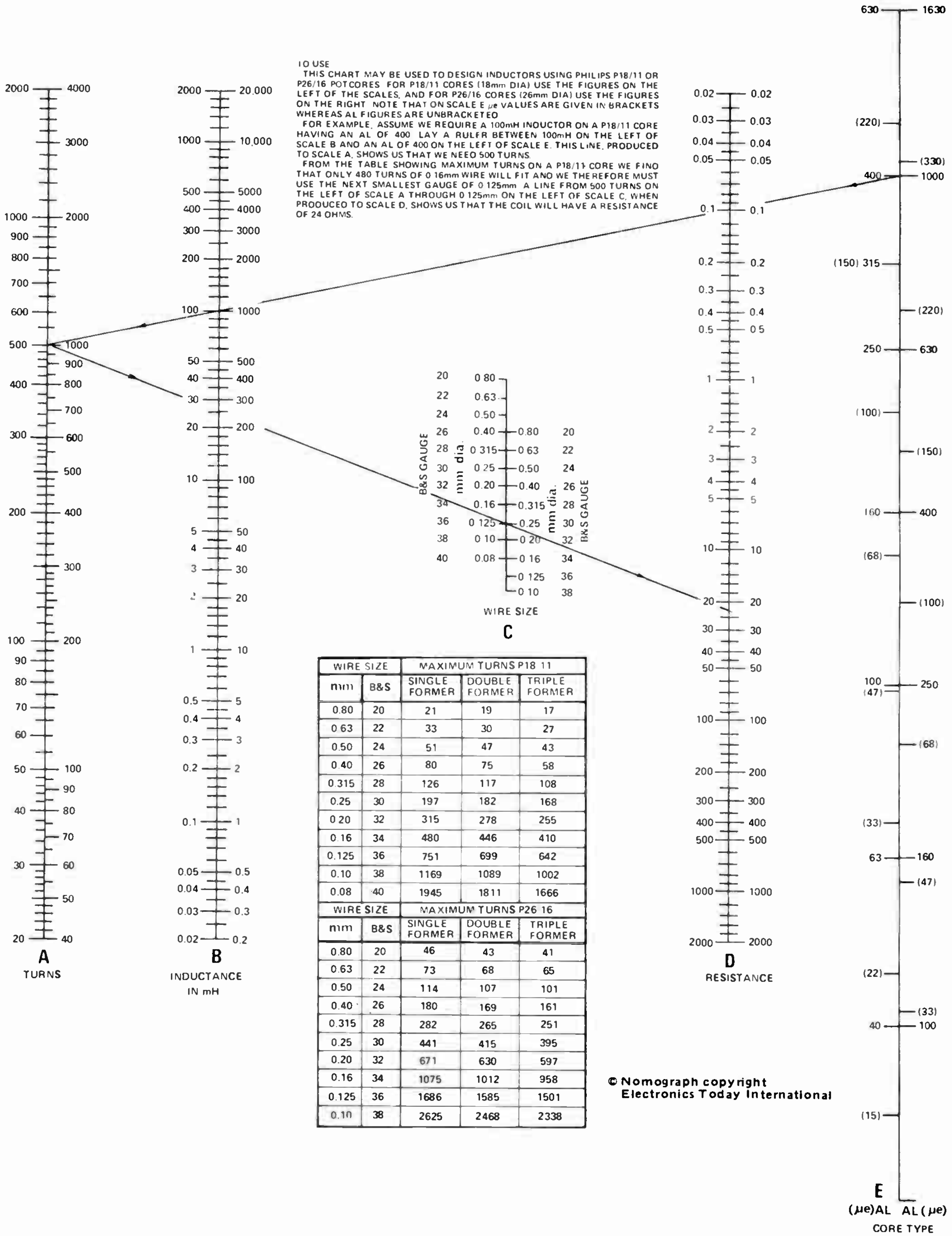
At higher ultrasonic and lower radio frequencies, additional factors of dielectric and skin-effect losses and parallel winding capacitance, all affect Q, making exact design difficult. Use of Litz wires, split section formers and small cores with low μe values will assist. ▶

TO USE

THIS CHART MAY BE USED TO DESIGN INDUCTORS USING PHILIPS P18/11 OR P26/16 POTCORES. FOR P18/11 CORES (18mm DIA) USE THE FIGURES ON THE LEFT OF THE SCALES, AND FOR P26/16 CORES (26mm DIA) USE THE FIGURES ON THE RIGHT. NOTE THAT ON SCALE E μe VALUES ARE GIVEN IN BRACKETS WHEREAS AL FIGURES ARE UNBRACKETED.

FOR EXAMPLE, ASSUME WE REQUIRE A 100mH INDUCTOR ON A P18/11 CORE HAVING AN AL OF 400. LAY A RULER BETWEEN 100mH ON THE LEFT OF SCALE B AND AN AL OF 400 ON THE LEFT OF SCALE E. THIS LINE, PRODUCED TO SCALE A, SHOWS US THAT WE NEED 500 TURNS.

FROM THE TABLE SHOWING MAXIMUM TURNS ON A P18/11 CORE WE FIND THAT ONLY 480 TURNS OF 0.16mm WIRE WILL FIT AND WE THEREFORE MUST USE THE NEXT SMALLEST GAUGE OF 0.125mm. A LINE FROM 500 TURNS ON THE LEFT OF SCALE A THROUGH 0.125mm ON THE LEFT OF SCALE C, WHEN PRODUCED TO SCALE D, SHOWS US THAT THE COIL WILL HAVE A RESISTANCE OF 24 OHMS.



WIRE SIZE		MAXIMUM TURNS P18 11		
mm	B&S	SINGLE FORMER	DOUBLE FORMER	TRIPLE FORMER
0.80	20	21	19	17
0.63	22	33	30	27
0.50	24	51	47	43
0.40	26	80	75	58
0.315	28	126	117	108
0.25	30	197	182	168
0.20	32	315	278	255
0.16	34	480	446	410
0.125	36	751	699	642
0.10	38	1169	1089	1002
0.08	40	1945	1811	1666

WIRE SIZE		MAXIMUM TURNS P26 16		
mm	B&S	SINGLE FORMER	DOUBLE FORMER	TRIPLE FORMER
0.80	20	46	43	41
0.63	22	73	68	65
0.50	24	114	107	101
0.40	26	180	169	161
0.315	28	282	265	251
0.25	30	441	415	395
0.20	32	671	630	597
0.16	34	1075	1012	958
0.125	36	1686	1585	1501
0.10	38	2625	2468	2338

Table 2 P18/11 Potcores

A. PRE-ADJUSTED PAIRS WITH STANDARD μ_e VALUES						
catalogue number 4322.022 . . .	grade of ferroxcube	effective permeability (μ_e)	number of turns for 1 mH α	tolerance on inductance %	adjustor type 4322.021	adjustor colour
28030	3B7	33	98.2	±1	30780	green
28040	3B7	47	82.3	±1	30800	red
2805C	3B7	68	68.4	±1	30980	white
28060	3B7	100	56.4	±1.5	30980	white
28070	3B7	150	46.1	±2	30810	brown
28080	3B7	220	38.1	±3	30810	brown
28090	3B7	330	31.0	±3	31090	grey
28230	3H1	33	98.2	±1	30780	green
28240	3H1	47	82.3	±1	30800	red
28250	3H1	68	68.4	±1	30980	white
28260	3H1	100	56.4	±1.5	30980	white
28270	3H1	150	46.1	±2	30810	brown
28280	3H1	220	38.1	±3	30810	brown
28290	3H1	330	31.0	±3	31090	grey
28430	3D3	33	98.2	±1	30780	green
28440	3D3	47	82.3	±1	30800	red
28450	3D3	68	68.4	±1	30980	white
28810	4C6	15	146	±1	30780	green
28830	4C6	33	98.2	±1	30790	yellow
08000	3B7	1910.0	12.9	±25	–	–
08200	3H1	1910.0	12.9	±25	–	–
08400	3D3	730.0	20.8	±25	–	–

$N = \alpha \sqrt{L}$ (L in 10^{-3} H)

B. PRE-ADJUSTED PAIRS WITH STANDARD AL FACTORS						
catalogue number 4322.022 . . .	grade of ferroxcube	AL factor	corres-ponding μ_e value	tolerance on inductance %	adjustor type 4322.021	adjustor colour
29030	3B7	63	20	±1	30780	green
29040	3B6	100	31.8	±1	30780	green
29050	3B7	160	51	±1	30800	red
29060	3B7	250	79.5	±1	30980	white
29070	3B7	315	100.2	±1.5	30980	white
29080	3B7	400	127	±2	30810	brown
29100	3B7	630	200	±3	30810	brown
29110	3B7	1000	318	±3	31090	grey
29120	3B7	1600	510	±3	31090	grey
29230	3H1	63	20	±1	30780	green
29240	3H1	100	31.8	±1	30780	green
29280	3H1	160	51	±1	30800	red
29260	3H1	250	79.5	±1	30980	white
29270	3H1	315	100.2	±1.5	30980	white
29280	3H1	400	127	±2	30810	brown
29300	3H1	630	200	±3	30810	brown
29310	3H1	1000	318	±3	31090	grey
29320	3H1	1600	510	±3	31090	grey
29430	3D3	63	20	±1	30780	green
29440	3D3	100	31.8	±1	30780	green
29450	3D3	160	51	±1	30800	red
29460	3D3	250	79.5	±1	30980	white
29830	4C6	63	20	±1	30780	green
29840	4C6	100	31.8	±1	30790	yellow

$L = N^2 AL$ (10^{-9} H)

C. COILFORMERS	
catalogue number	number of sections
4322.021.30330	1
4322.021.30340	2
4322.021.30350	3
4322.021.30130	1 with pins
4302.021.20030	1 with pins

D. MOUNTING PARTS	
catalogue number	description
4322.021.30550	container
4322.021.30660	spring
4322.021.30470	tag plate
4322.021.30710	nut
4322.021.30720	bush
4302.021.20020	clip

Inductance

The tolerance given on inductance, in Tables 2 and 3 is obtained when using the specified core, and a wire size that will completely fill the former close layer wound. Due to slight changes in wire diameter and different methods of winding, the exact number of turns accommodated may vary by ±10%.

Table 3 P26/16 Potcores

A. PRE-ADJUSTED PAIRS WITH STANDARD μ_e VALUES						
catalogue number 4322.022	grade of ferroxcube	effective permeability (μ_e)	number of turns $\propto$	tolerance on inductance %	adjustor type number 4322,021 ...	adjustor colour
24030	3B7	33	120	±1	30760	green
24040	3B7	47	100.5	±1	30770	red
24050	3B7	68	83.6	±1	30960	yellow
24060	3B7	100	68.9	±1.5	30970	white
24070	3B7	150	56.3	±2	30730	brown
24080	3B7	220	46.5	±3	31080	grey
24230	3H1	33	120	±1	30760	green
24240	3H1	100.5	100.5	±1	30770	red
24250	3H1	68	83.6	±1	30960	yellow
24260	3H1	100	68.9	±1.5	30970	white
24270	3H1	150	56.3	±2	30730	brown
24280	3H1	220	46.5	±3	31080	grey
24430	3D3	33	120	±1	30760	green
24440	3D3	47	100.5	±1	30770	red
24450	3D3	68	83.6	±3	30960	yellow
24810	4C6	15	178	±1	30760	green
24820	4C6	22	147	±1	30770	red
24830	4C6	33	120	±1	30970	white
04000	3B7	1750	16.5	±25	—	—
04200	3H1	1750	16.5	±25		
04000	3D3	705	25.9	±25		

$N = \propto \sqrt{L}$ (L in 10^{-3} H)

B. PRE-ADJUSTED PAIRS WITH STANDARD AL FACTORS						
catalogue number 4322.022	grade of ferroxcube	AL factor	corres-ponding μ_e value	tolerance on inductance $\propto$	adjustor type 4322.021 ...	adjustor colour
25030	3B7	63	30	±1	30760	green
25040	3B7	100	47.5	±1	30770	red
25050	3B7	160	76	±1	30960	yellow
25050	3B7	250	119	±1.5	30970	white
25070	3B7	315	149	±2	30730	brown
25080	3B7	400	190	±2	31080	grey
25100	3B7	630	298	±3	31080	grey
25230	3H1	63	30	±1	30760	green
25240	3H1	100	47.5	±1	30770	red
25250	3H1	160	76	±1	30960	yellow
25260	3H1	250	119	±1.5	30970	white
25270	3H1	315	149	±2	30730	brown
25280	3H1	400	190	±2	31080	grey
25300	3H1	630	298	±3	31080	grey
25420	3D3	40	19.0	±1	30760	green
25430	3D3	63	30	±1	30760	green
25440	3D3	100	47.5	±1	30770	red
25450	3D3	160	76	±1	30960	yellow
25810	4C6	25	11.9	±1	30760	green
25820	4C6	40	19.0	±1	30770	red
25830	4C6	63	30	±1	30970	white

Hence it is safer, when winding experimental coils, to only try and fit 90% of the turns indicated in the maximum number of turns tables.

If the former is only partly filled, errors up to 4% may occur with the lower μe cores. However the use of an adjustor will allow a +10% adjustment range which is generally sufficient to cope with tolerances found in practical circuits.

When optimum stability is required the type of adjustor that matches a certain core should be used. If it is desired to widen the adjustable range, at the possible expense of stability, an adjustor indicated for a potcore with a high μe value may be used with a potcore of low μe value. ●

C. COIL FORMERS	
catalogue number	number of sections
4322.02130270	1
4322.021.30280	2
4322.021.30290	3
4322.021.30090	1 with pins
4302.021.20010	1 with pins

C. MOUNTING PARTS	
catalogue number	description
4322.021.30530	container
4322.021.30640	spring
4322.021.30450	tag plate
4322.021.30710	nut
4322.021.30720	bush
4302.021.20000	clip

Design data for this article has been derived from the Philips publication "Ferroxcube Potcores", 1972. Nomograph copyright — Electronics Today International. This article was originally published in the October 1974 issue of ETI, Australian edition.

Precision temperature measurement and control using the LM335

Brian Dance

Measuring temperature for its own sake or in a control application requires a suitable 'transducer' to provide an accurately known relationship between temperature and output. That's just what the LM335 does.

THERE ARE MANY ways of measuring temperature, but the familiar mercury-in-glass thermometer does have the disadvantage that it is not easily read and remote readout is impossible. The circular clock-face type of thermometers based on bimetallic strip in the form of a coil are very convenient for hanging on the wall in the home or greenhouse, but have a very limited accuracy. Electronic thermometers providing a very accurate digital indication of temperature are very convenient, although the commercially available types are necessarily moderately expensive.

This article describes the use of a device specially developed by National Semiconductor for the precision measurement of temperature which can be used in circuits whose output is usually fed to a digital voltmeter so that a digital indication of temperature can be obtained.

The LM335

The LM335 is an integrated circuit temperature sensor for use over the range 0° C to +100° C. It is available in economical plastic packaging with the connections shown in Figure 1, although a similar device is available in a TO-46 metal transistor type package with the connections of Figure 2.

The LM335 is a relatively economical device, but the LM235 is a similar product with the same internal circuitry designed for use over the −25° C to +100° C range and the LM135 can operate over the military temperature range of −53° C to +150° C; these last two devices have narrower tolerance than the LM335 specifications. Suffix 'A' versions, such as the LM335A, are also manufactured with more closely specified characteristics. However, it will be assumed that readers will employ the most economical device in the range, the LM335, although the circuits can be used with any of the devices named.

Basically, the LM335 is operated in the same way as a zener diode, as shown in the circuit of Figure 3. The breakdown voltage (that is, the output voltage from this circuit) is directly proportional to the absolute temperature and is 10 mV/K over the specified working temperature range.

The value of R1 in Figure 3 determines the current flowing through the device, but as the dynamic impedance at 1 mA is typically 0.6 Ohm, the device can be operated over the current range of 400 μA to 5 mA with virtually no change in its performance. It should be noted that the absolute maximum forward or reverse current which may safely be passed through the device (even momentarily) is only 10 mA; higher currents may cause irreversible damage to the LM335.

At 25° C and a reverse current of 1 mA, the operating output voltage from the Figure 3 circuit is typically 2.98 V with minimum and maximum limits of 2.92 V and 3.04 V. The value chosen for R1 may be calculated for a current through the LM335 of 1 mA using the equation R1 = $(V+ - V_{out})/0.001$ which equals approximately (V+ - 3) kilohm.

Linear output

A particular advantage of the LM335 is the linear output provided by its circuit, unlike the output of most other temperature sensors which is not linearly related to temperature. Indeed, if the output voltage is plotted against temperature over the working range and the graph is extrapolated back to the absolute zero of temperature, the output read from the graph at the latter temperature will be zero.

Although the LM335 output from the Figure 3 circuit is within the limits stated a calibration connection is included on the chip. It is only necessary to connect a potentiometer across the LM335, as shown in Figure 4, and adjust this potentiometer to 2.982 V output when the device is at 25° C in order to obtain higher accuracy over the whole temperature range.

The single calibration temperature over the whole working range is possible because the output is accurately proportional to the absolute temperature with the extrapolated output falling to 0 V at the absolute zero of temperature. Variations from one LM335 to another are only in the slope of the voltage/temperature graph, so a slope calibration at one temperature corrects for all others. Thus, calibration is far easier than with a non-linear device such as a thermocouple.

Self heating

As with any temperature sensing system, any heat generated by the current passing through the sensing device will affect its temperature and hence the output voltage. The LM335 should therefore be operated at the lowest current which is adequate to drive its internal circuitry. When calculating the value of R1 allowance should be

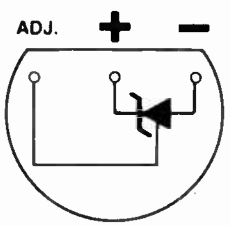


Figure 1. The TO-92 plastic encapsulated LM335.

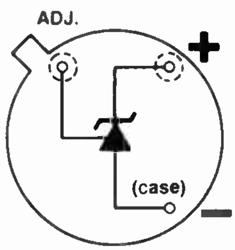
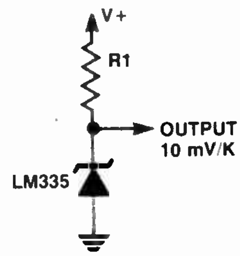


Figure 2. The TO-48 metal encapsulated LM335.

Figure 3. The basic LM332 circuit.



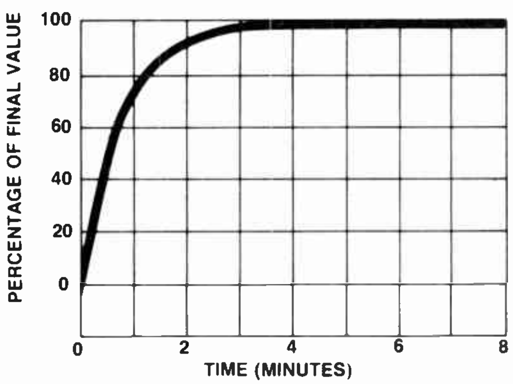


Figure 5. Response time of the LM335 to a temperature change in free air.

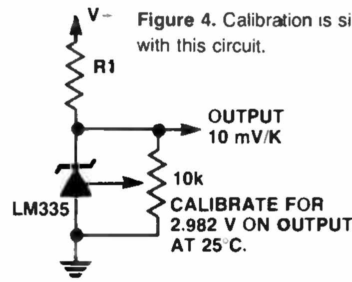


Figure 4. Calibration is simple with this circuit.

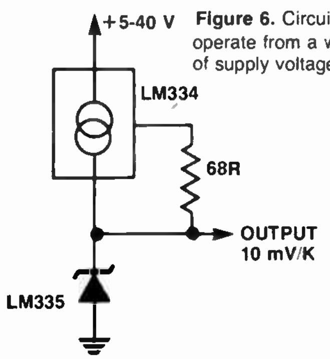
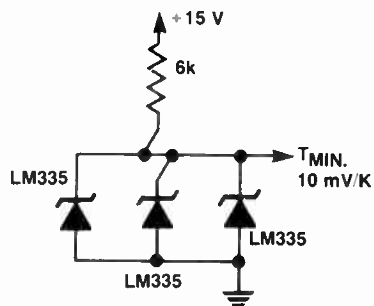


Figure 6. Circuit able to operate from a wide range of supply voltages.

Figure 7. This circuit records the minimum temperature of any of the devices.



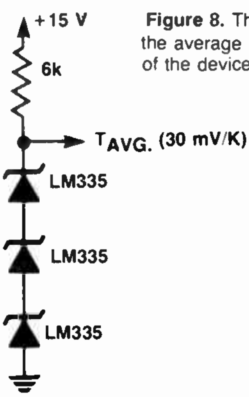


Figure 8. This circuit records the average temperature of the devices.

made for the current passing through any calibrating potentiometer in parallel with the device and for any output current. A current of about 400 μA is about the normal minimum.

If the sensor is used in a situation where the thermal resistance to the surroundings is constant, self-heating errors can be calibrated out, provided the device is operated with a constant current independent of temperature. Heating of the device will then be proportional to the zener voltage and to the absolute temperature; thus, the self-heating error is proportional to the absolute temperature and temperature scale linearity is preserved.

Performance

In a typical LM335 circuit which has not been calibrated, operating at 1 mA, the temperature error is 2° C (maximum 6° C) at 25° C or 4° C (maximum 9° C) over the whole working range. When calibrated the typical error at the temperature limits is 2° C. Non-linearity at 1 mA is typically 0.3° C over the range.

In still air the device requires about three minutes to reach its final temperature after a temperature change has occurred (Figure 5), the time constant being typically 80 seconds. In stirred oil the final temperature is reached within about three seconds (time constant typically one second). The device is stable to 0.2° C (typical) over 1000 hours, even at 125° C.

The dynamic impedance is less than one Ohm at frequencies up to more than 1 kHz (typical), but increases to 20 to 30 Ohms at 100 kHz.

Circuits

The circuits of Figures 3 and 4 are suitable for use when the supply voltage is reasonably constant. If wide variations in the supply voltage are expected to occur, the LM334 constant current device may be used with the external resistor to set the LM335 current at about 1 mA for all supply voltages. (Figure 6.)

If a number of LM335 ICs are connected in parallel, as in Figure 7, the output will correspond to that of the device which is at the minimum temperature. Thus a minimum indication of the temperatures at three locations is easily obtained.

Similarly, a number LM335 devices may be connected in series, as in Figure 8, in which case the output will represent the average temperature of the devices, but will be increased by a factor equal to the number of devices used.

Centigrade Thermometer

The circuits discussed previously are basic ones which provide an output voltage directly proportional to the absolute temperature, but this is not very consistent for feeding to a digital voltmeter to produce a reading directly in °C. The additional operational amplifier circuit of Figure 9 is required for this purpose.

In this circuit the LM336 provides a precise 5 V reference voltage to pin 3 of the LM308 operational amplifier. The negative feedback to pin 2 is adjusted by means of the 2k potentiometer so that the output of the amplifier is at a potential of 2.73 V. The voltage difference between this output and that from the LM335 circuit is then a measure of the Centigrade ▶

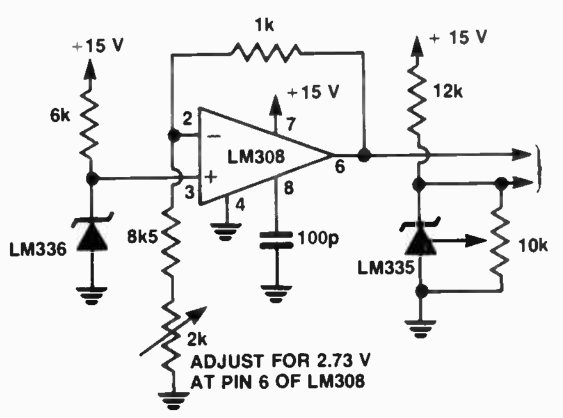


Figure 9. A centigrade thermometer circuit.

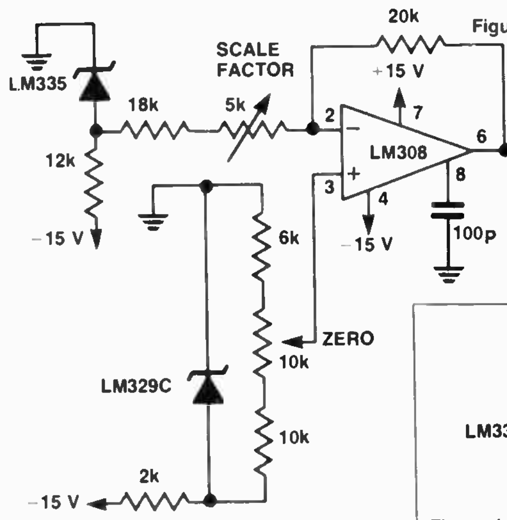


Figure 10. A centigrade thermometer with output referenced to ground

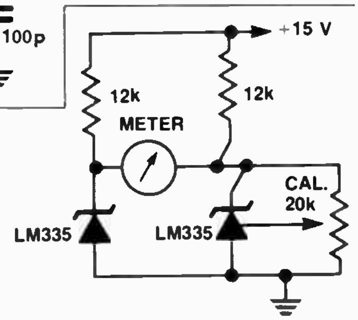


Figure 11. A differential temperature sensor.

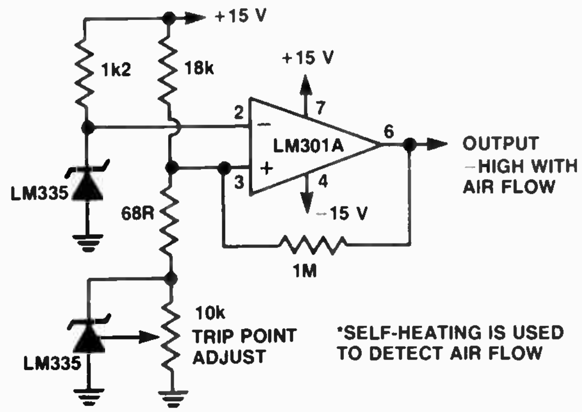


Figure 12. A differential temperature sensor with amplifier gain of ten.

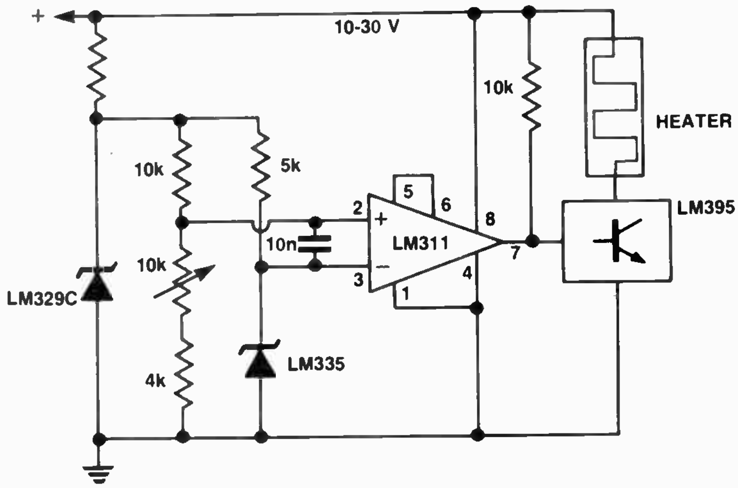


Figure 13. A thermostatic heater controller.

temperature: the 2.73 V reference effectively subtracts 273° C from the absolute temperature indicated by the output from the LM335 circuit to leave the Centigrade temperatures to be displayed by a digital voltmeter set to an appropriate scale.

Neither of the outputs from the circuit of Figure 9 are at ground potential. The slightly more complex circuit of Figure 10 provides an output of 10 mV/° C referred to ground. It employs an LM329C 6.9 V precision reference voltage device to provide a variable preset voltage to the non-inverting input of the LM308 operational amplifier. The latter takes its inverting input to a feedback network involving the LM335 temperature sensing device.

Differential sensors

Two LM335 devices in different positions can be used in the simple circuit of Figure 11 to measure the temperature difference between the two positions. Only one calibration control is required to give a zero difference when the two devices are at the same temperature.

In Figure 12 an operational amplifier is used to compare the outputs of two LM335 devices connected as in Figure 11, but the negative feedback circuit is arranged to provide a gain of ten so that the output from the amplifier is the Centigrade temperature difference in 100 mV/° C.

Temperature controller

A simple temperature control circuit which adjusts the current through a heater to maintain the temperature at some constant desired value is shown in Figure 13. The LM329C provides a precision 6.9 V reference, the fraction of this reference voltage which is tapped off and fed to the non-inverting input of the LM311 being adjusted by the temperature setting potentiometer.

If the temperature of the LM335 is high enough for the voltage from it (which is fed to the inverting input of the LM311) to exceed that of the non-inverting input, the output of the amplifier will be low so that the LM395 passes only a very small current through the heater. If the LM335 temperature falls, the LM311 output rises and switches on the LM395 so that current passes through the heater. The LM395 is actually an IC which behaves like a very high gain power transistor.

Air flow detector

In the circuit of Figure 14, a fairly high current is passed through the upper LM335 so that the device becomes warm. If air flows fairly quickly past this device, it will be cooled and its output voltage will fall. As this voltage is connected to the inverting input of the LM301A device, the output of the latter will become 'high' when such a fast air flow occurs, the lower LM335 (not in the air flow) is used to provide a comparison voltage by keeping the ambient temperature around the two LM335 devices the same.

Fast NiCad charger

Nickel cadmium cells can be fast charged only if precautions are taken to ensure that the temperature of the cells does not rise above a permissible limit. In the circuit of Figure 15, the LM335 diode D2 is placed in close thermal contact with the Nickel Cadmium cells being charged. If the temperature of the cells rises, the output from D2 rises and, as this output is

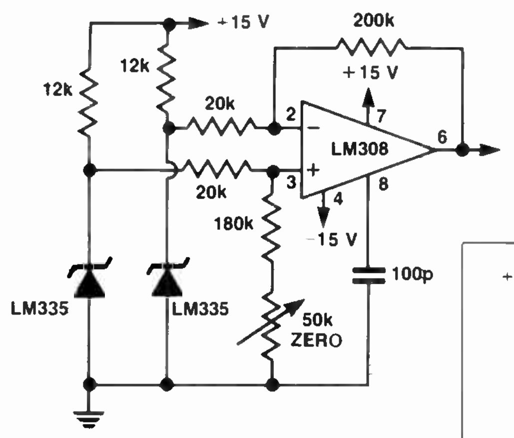


Figure 14. An air-flow detector.

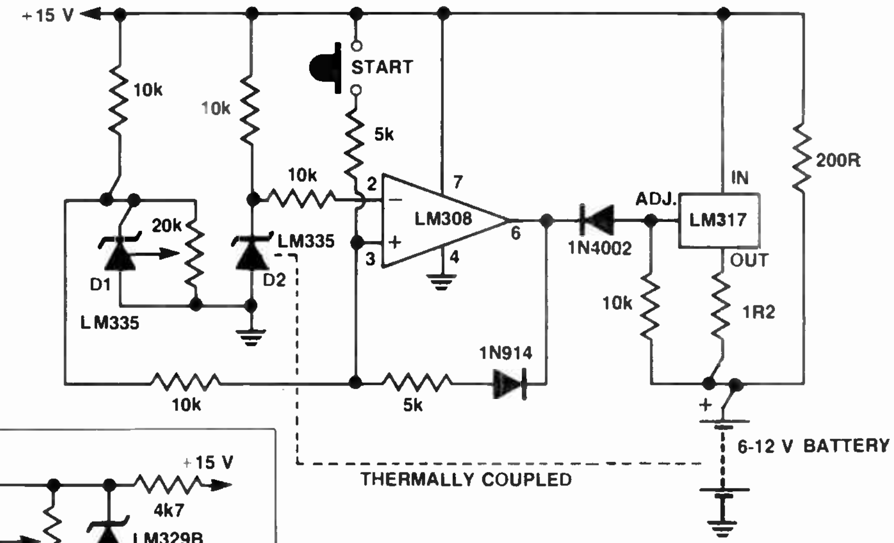
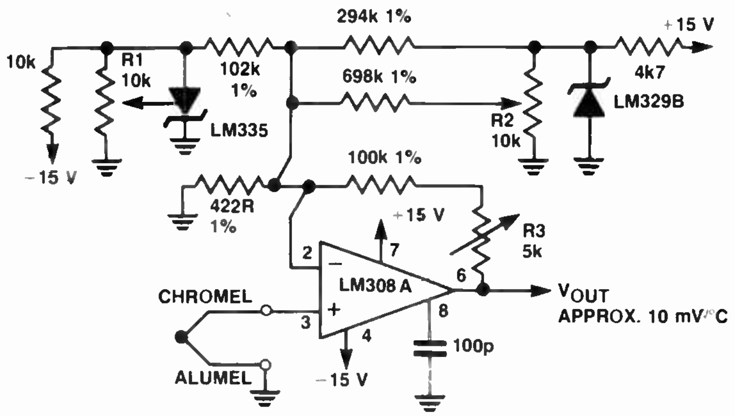


Figure 15. A fast charging circuit for Nickel Cadmium batteries with temperature monitoring and automatic switch-off in the event of over heating.

Figure 16. A centigrade calibrated thermocouple electronic thermometer.



fed to the inverting input of the LM308, the output of this operational amplifier falls.

Current passing from the 'ADJ' terminal of the LM317 regulator to the LM308 causes the potential at this terminal to fall so that the regulator no longer passes the charging current to the nickel cadmium cells. The non-inverting input potential is derived from the other LM335 which is at the ambient temperature. Thus the temperature of the cells is compared with the ambient temperature, as required. The calibration potentiometer across D1 may be adjusted so that the voltage across this LM335 is 50 mV greater than that across D2. The charging will then be terminated when the temperature of the cells rises by 5° C above ambient.

Thermocouple application

Thermocouples are much used for making temperature measurements over a much wider range than is possible with the LM335, partly because they are very cheap and simple. Although thermocouples can be used for measuring temperatures of up to some thousands of degrees Centigrade using a junction of two different materials, a cold reference junction is required, often an ice bath, except when differential measurements are being made.

Rather than use an ice bath it is often more convenient to employ a technique known as cold junction compensation in which a compensating voltage is added to the output of the thermocouple so that the reference junction potential seems to be at 0° C, although it is actually at another temperature. The added voltage can be made proportional to temperature with the same constant of proportionality as the thermocouple so that changes in ambient temperature have no effect on the output voltage.

The LM335 temperature sensor is very suitable for use in the cold junction compensating circuit owing to its very linear voltage/temperature characteristics. In addition, as the LM335 voltage extrapolates to zero at the absolute zero of temperature, the temperature coefficient of the compensation circuit can be adjusted to room temperature without any temperature cycling.

A thermocouple thermometer calibrated in degrees centigrade is shown in Figure 16. The thermocouple reference junction should be terminated in close proximity to the LM335 so that their temperatures do not differ appreciably. Initially a signal should be applied in place of the thermocouple and R3 adjusted for a gain of 245.7. The non-inverting input of the LM308A should be shorted to ground and R1 adjusted so that the output voltage is 2.982 V at 25° C. The short should now be removed from the non-inverting input and R2 adjusted for an output of 246 mV at 25°C. The thermocouple connections should now be replaced.

Electronic thermometers of this general type can provide a 10 mV/° C output over a 0° C to 1300° C range, but it is important to use good quality cermet trimmers and stable components.

Further thermocouple circuits together with practical information on their construction using LM335 cold junction compensation is available in the *National Semiconductor Linear Applications Handbook*, as Application Note AN-225, April 1979.

Optocoupler devices and their applications

Ever wanted to control one circuit with another without having any intermediate electrical connection? Devices that provide coupling via a beam of infra-red light are called 'optocouplers' and they're just perfect for the job. Here's a run down on a host of popular optocouplers and how to use them in practical circuits.

Brian Dance

THE ELECTRONIC circuit designer is often faced with the problem of providing a high degree of isolation between two circuits which must nevertheless be able to pass alternating signals from one part of the circuit to the other.

For example, one may wish to have one part of the circuit completely isolated from the mains, yet use signals from this part of the circuit to control the flow of the mains current through a load. Another example occurs in patient monitoring equipment where the small voltages developed by the beating of a heart can be coupled into mains powered equipment without any danger of the equipment causing a current to flow through the heart.

Optocouplers

Optocouplers use a beam of infra-red radiation (or occasionally, visible light) to convey the signal from one part of the circuit to the other without any electrical connection whatsoever between the two parts. They are sometimes known as photon-coupled devices or as opto-isolators. They may be employed to replace conventional relays when a fast response is required or when sparking at relay contacts must be avoided in an explosive atmosphere.

An optocoupling device consists of an infra-red emitting device or other lamp on its input side and some form of detector for the radiation on the output side, both the emitter and detector being in a light-tight enclosure. The silicon detector itself may be a phototransistor, a photo-Darlington device, an opto-triggered triac or even a field effect phototransistor.

No matter which of these device types is employed, the silicon detector has its maximum sensitivity at a wavelength quite near to that at which the gallium arsenide device emits with its maximum intensity. In other words, the devices are spectrally well matched so that a small emitter device current can produce a reasonably large response in the detecting device.

Types

A very large number of types of optocoupler have been marketed with the electrical characteristics of both the emitter and the detector having to be specified in every type. Rather than involve readers with a mass of type numbers, this article will concentrate on a limited number of readily-available devices.

The 4N26, 4N28 and MCT2 devices are examples of those using a phototransistor as a detector, the 4N33 has a Darlington output stage, the 6N139 (equivalent to the MCC671) has a 'split-Darlington' fast output device, the MCT6 is a dual device and the MOC-3020 has a triac output for 240 V mains supplies.

Dual-in-line

Although some optocoupled devices are fabricated in circular metal packages, the most common types, including those listed above, are produced in dual-in-line (DIL) packages with a typical construction like that shown in Figure 1. The emitter and detector

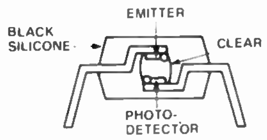


Figure 1. Cross-section through an optocoupler.

are placed fairly close together with a clear insulating material between them. The black silicon body of the device prevents stray radiation from falling on the detector. A circuit symbol is shown in Figure 2.

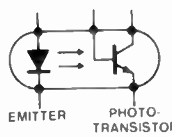


Figure 2. Symbol for an optocoupler having a transistor output stage.

In most DIL devices the radiating emitter is connected to pins on one side of the device, while the detector is connected to pins on the other side. This arrangement provides the maximum possible electrical isolation between the input and output circuits. Many of the simpler optocoupled devices differ from most other dual-in-line devices in that they have a total of only six connecting pins.

The basic internal circuitry of the devices under discussion is shown in Figures 3 to 7 inclusive. The three devices 4N26, 4N28 and MCT2 with a single phototransistor output all have the connections shown in Figure 3. The dual device type MCT6 is housed in the 8-pin package of Figure 4 so that the additional pins required are available.

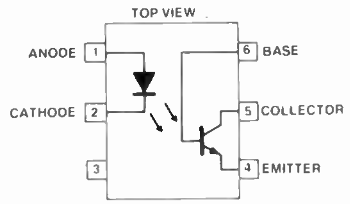


Figure 3. Pinout for the 4N26, 4N28 and MCT2 devices.

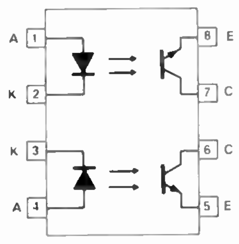


Figure 4. Pinout of the MCT6 dual optocoupler device.

DEVICE TYPE	4N26	4N28	MCT2	MCT6 (dual)	4N33	6N139	MOC-3020
Output device	←	Phototransistor(s)		→	← Darlington	→	Triac
CTR (%) Min.	20	10	—	5	—	400	—
Typ.	50	30	—	50	500	—	—
Isolation (kV)	1.5	0.5	1.5	1.5 (0.5 between channels)	1.5	3	7.5 (max 5 sec)
Isolation resistance (Typical ohms)	10^{11}	10^{11}	10^{12}	10^{12}	10^{11}	10^{12}	—
Isolation capacitance at 1 MHz (pF)	1.3	1.3	0.5	0.5	0.8	0.6	—
Maximum emitter current (mA)	80	80	60	60 per emitter	80	20	50
Typical emitter voltage at 50 mA	1.2	1.2	—	—	1.2	1.5 (at 1.6 mA)	—
Maximum reverse input voltage (V)	3	3	3	3	5	5	3
Input capacitance (pF)	150	150	250	—	150	—	—
Maximum power (mW) Total	250	250	250	400	250	—	300
Input	150	150	200	100	150	35	
Output	150	150	200	150	150	100	
Output transistor:							
BV_{CEO} (min.)	30	30	85	85	30	—	
BV_{CBO} (min.)	70	70	165	—	50	—	
BV_{ECO} (min.)	7	7	14	13	5	—	
h_{FE} (typ.)	250	250	60	—	5000	—	
I_{CEO} (nA typ.)	50	100	50	50	100	1000	
$V_{CE(SAT)}$ (typ.)	0.2	0.2	0.24	0.24	1 0	—	
Typical I_c for I_F = 10 mA	5	3	—	—	—	—	
Typical bandwidth (kHz)	300	300	150	150	30	—	
Package	Figure 3	Figure 3	Figure 3	Figure 4	Figure 5	Figure 6	Figure 7

MOC-3020:
$V_{DRM} = V_{RRM}$ = 400 V
Trigger current 5 mA (typ.), 20 mA (max.)
Holding current 100 uA (typ.) at V_{AK} = 3 V
Maximum output current 100 mA RMS

TABLE 1. Basic data on the types of optocouplers discussed.

The 4N33 with its high-gain photo-Darlington output device is encapsulated in a 6-pin package with the same type of connections as the phototransistor output types of Figure 3; except for the performance differences, these devices are pin-for-pin replaceable.

The 6N139 'split-Darlington' output device has its output transistor base brought out to a separate pin, so the 8-pin dual-in-line package of Figure 6 is employed; this enables the input diode connections to be kept on the opposite side to all of the output connections.

Finally, the MOC-3020 with its triac output stage, is housed in a 6-pin dual-in-line package with the connections shown in Figure 7.

The basic parameters of these devices are listed in Table 1, but it cannot be over-emphasised that some of these values apply only under certain operating conditions stated on the data sheet which cannot all be shown in a table of a reasonable size.

It can be seen that most of the specifications required for the MOC-3020 differ from those of the other devices in their nature owing to the fact that the output triac must be specified in a different way to transistors and Darlingtons.

Which type?

If one wishes to use an input signal to control alternating current from the mains in a load, the MOC-3020 will generally be the best device from those under discussion. This optocoupler will be discussed separately from the others.

If one has to design a circuit which requires two separate control coupling systems, this can be done using the dual MCT6 device provided that phototransistor outputs are suitable for the particular application concerned. Indeed, two of these MCT6 devices can be inserted into a 16-pin dual-in-line IC socket so that one has a quadruple coupling system. (Quad devices in a single package are manufactured, but are not so common as the types under discussion.)

This leaves us with a choice, in the case of single devices, of those using a phototransistor or those employing a photo-Darlington output stage. The types using a phototransistor are most commonly employed, since they provide a fast response and can usually handle input signals with frequencies of over 100 kHz (see Table 1).

Photo-Darlington output devices provide a higher gain, but the bandwidth (or maximum usable signal frequency) is about an order of magnitude less than devices which use a simple phototransistor output; in addition, devices using a photo-Darlington output stage may be priced some 50% higher than those employing a phototransistor output, although this is not always the case.

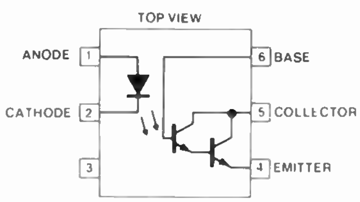


Figure 5. Pinout of the 4N33, an optocoupler having a photo-Darlington output stage.

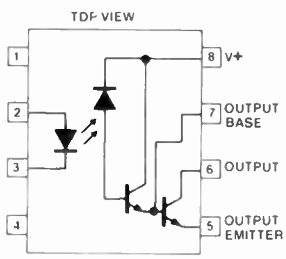


Figure 6. Pinout for the 6N139 (or MCC671) device which has a separate photodiode for maximum speed and a Darlington output for high gain.

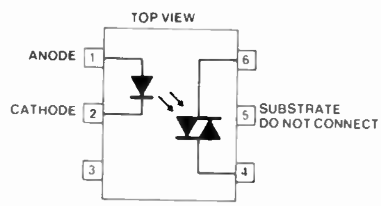


Figure 7. Pinout of the MCO-3020, which has a triac output stage.

The single devices of Figures 3 and 5 (but not the dual device of Figure 4) have the transistor base connected to a separate pin so that suitable circuitry may be used to trade-off gain in order to obtain a better high frequency response. The maximum usable frequency will be obtained when the output phototransistor is connected as a photodiode using only the base and collector connections of Figure 3, but a relatively large input current will then be needed to produce a small output current; the CTR value may be under 0.1%.

Apart from the more limited response speed of a device with a photo-Darlington output stage, it can be seen from Table 1 that the saturation voltage (under high input conditions) is much greater for the photo-Darlington device than for a simple phototransistor. Both the speed of response and the saturation differences are inherent properties of photo-Darlington devices and are not limited to optocouplers.

CTR

In order to understand some of the figures quoted in Table 1, we must first examine the ways in which certain device parameters are specified. The user need not consider any of the internal optical design points, since the manufacturer takes care of such considerations when he is designing the devices concerned.

Optocouplers are supplied as sealed units, although opto-interrupter modules are also manufactured in which there is a slot between the emitting diode and the detector so that a metal vane passing through the slot can interrupt the beam; such opto-interrupters can, for example, be used in car ignition timing systems.

One of the most important parameters of an optocoupled device is its *current transfer ratio* (CTR) which is the ratio of the output current to the input current under certain conditions specified by the manufacturer; it is usually expressed as a percentage and, broadly speaking, may be considered as the 'gain' of the device. It may be noted that devices with a triac or a thyristor output do not have a CTR value.

It can be seen from Table 1 that typical values of the CTR in the case of devices which have a simple phototransistor output stage is of the order of 50% — which means the collector current in the output phototransistor will be about half that to the input diode emitter.

The minimum value in a device of any specified type may be considerably less than that of the typical value. However, in the case of devices with photo-Darlington outputs, a CTR value of 500% is more common — which means the output current is five times the input current.

In some special devices a short light pipe is used to carry the radiation from the emitter to the detector, inevitably with some loss, so the CTR value may be reduced in such devices which may be able to withstand a much higher voltage between their input and output sides.

Unfortunately, the CTR does not have a constant value but varies widely with the diode input current and with the device temperature. Figure 8 shows the typical variation of the CTR value of the MCT2 device (which has a simple phototransistor output stage) with the forward input current passed through the emitter diode.

Each curve is for a different MCT2 device, the wide spread being due to variations in the phototransistor gain, the emitter efficiency and the coupling efficiency between the two internal components. The percentage values quoted on each curve are those for a 10 mA input current.

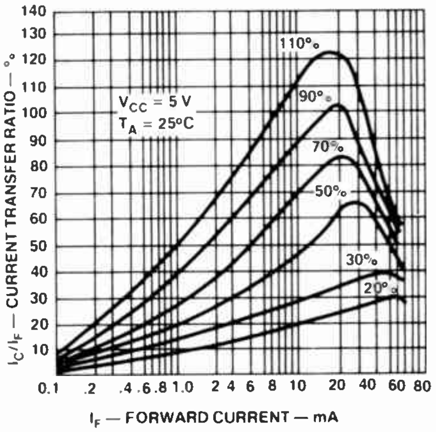


Figure 8. Variation of the current transfer ratio (CTR) with forward current in typical MCT2 devices.

The CTR value of a 4N26 or 4N28 can vary by a factor of about 2.5 between high temperatures (where it is relatively low) and very low temperatures, while devices with Darlington outputs may show variations of double this factor between temperature extremes. Rather smaller variations are more commonly found.

Isolation

Manufacturers of optocoupled device specify a maximum voltage which may be safely applied between the input and output sides of the device. In most devices this is in the range 500 to 8000 V, depending on the device type, but special types can be obtained for higher voltage isolation.

The resistance between the input and output sides of a typical device is often around 10^{11} to 10^{12} ohm. Although this seems very high, if a potential of a few kilovolts is applied across the device, a current of somewhat under 100 nA can flow. This is comparable with the current through the output of a high gain device when the input current through the emitter is under 1 mA.

If an optocoupler fails under a high applied voltage between its input and output sides, a short circuit will normally develop as a track is formed between the emitting and detecting devices. The problem can be reduced by the use of suitable current limiting resistors or protective devices in either the input or output circuit.

The stray capacitance between the input and output circuit of an optocoupler is typically of the order of 1 pF (Table 1). It can provide some unwanted coupling in circuits designed to be able to operate at high speeds, especially when inductive loads are being switched.

The emitter

The emitting diode will have a maximum continuous current rating, normally some tens of milliamps as indicated in Table 1. In some devices, pulsed currents above the maximum continuous current are permissable.

A maximum value is also imposed on the reverse voltage which may appear across the emitter diode. The application of a higher reverse voltage can cause it to breakdown and perhaps pass a destructive current; however, this problem is easily avoided by connecting an external diode across the emitter diode as shown in Figure 9.

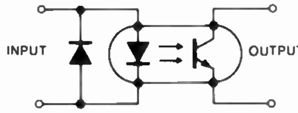


Figure 9. If a reverse voltage is likely to appear across the optocoupler emitter an external diode can be used to 'clip' it.

Although gallium arsenide diodes have been the main type used in optocouplers, there is an increasing trend to employ gallium-aluminium-arsenide types, since the latter not only emit photons more efficiently, but also provide a slightly better spectral match to the silicon detector. Thus an appreciable increase in the CTR value can be obtained.

In many optocouplers one must be careful to observe not only the total power dissipated in the complete package, but also the power dissipated in the separate input and output devices, as indicated in Table 1.

The detector

As with any other phototransistor or photo-Darlington, there is a certain value quoted for the maximum voltage which may be applied between the collector and the emitter with the base unconnected without risk of the device undergoing breakdown; this is BV_{CEO}. Similarly, values may be quoted for BV_{CBO} and BV_{ECO}.

A maximum collector current may also be quoted together with a maximum collector leakage current with base unconnected, I_{CEO}, under specified conditions.

The characteristics of the detector determine the speed of response and the bandwidth, since the emitting diodes are fast. The response time can be reduced by the use of a smaller value of load resistor, but many manufacturers quote rise and fall times and bandwidths with load resistors which are so small that the circuit would have an inadequate gain for most applications.

The response speed of an optocoupler can be improved by using the circuit of Figure 10 in which the collector load is effectively reduced to a very low value by the virtual earth input impedance of the operational amplifier.

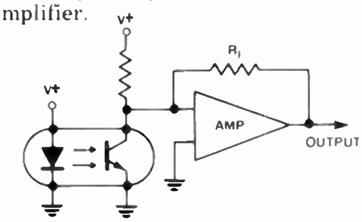


Figure 10. Response speed may be increased by the use of the virtual earth input of an op-amp.

An even simpler way of obtaining a faster response at the expense of a reduced value of the CTR involves connecting a resistor, between the base and emitter of the output transistor. As the value of this resistor is reduced, the response becomes faster until in the limit, when the resistor is a short circuit, one is using the detector as a photodiode.

If one expects to be working with a very small input current, one might expect the use of a high gain device with a photo-Darlington output would be ideal. This is not necessarily true, since the overall efficiency can fall at such currents to the point where a device with a phototransistor would be better.

APPLICATIONS

Optocoupling devices can be employed to replace relays and pulse transformers in a wide variety of applications in which high isolation may be desirable or essential. They provide fast signal transfer with excellent noise immunity. They are suitable for interfacing with TTL and CMOS circuits and can also be used for analogue signal coupling.

Circuits designed for use with single phototransistor output optocoupled devices can generally employ the 4N26, 4N28 or MCT2, but note should be made of the individual differences listed in Table 1.

For example, the 4N28 is limited to applications in which the voltage across the device does not exceed 500 V, while when the other devices are selected, it may be as great as 1.5 kV.

The phototransistors in the MCT2 and in the dual MCT6 outputs are much higher voltage devices than those used in the 4N26 and 4N28.

The bandwidth of the 4N26 and 4N28 is typically greater than that of the other two types, but so is the isolation capacitance between the input and output. However, these points are not likely to be of any great importance in most applications.

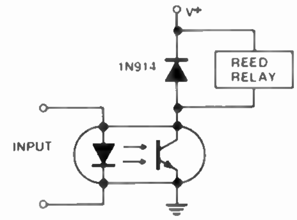


Figure 11. Using an optocoupler to isolate a reed relay.

Relay control

The simple circuit of Figure 11 shows how a small input current may be employed to control a reed relay. The inductive back-emf from the relay coil formed when the current ceases to flow through it is by-passed by the 1N914 diode so that this relatively high voltage pulse cannot damage the output transistor of the optocoupler.

The supply voltage used, V^+, should have a value about equal to the voltage required by the relay, but should not exceed the V_{CEO} value of Table 1 for the optocoupler used.

Although the use of a reed relay is suggested so that the output current of the optocoupling device is kept quite small, other types of small relay can be controlled with careful circuit design. Obviously this type of circuit provides better isolation than many types of relay.

The circuit can easily be modified so that the relay does not close until the input has been applied for a short time. One merely connects a capacitor across the input diode and feeds this diode through a series resistor. The delay time before the relay closes will be dependent on the time taken for the capacitor to charge through the series resistor.

Isolated audio

The circuit of Figure 12 shows how an audio output completely isolated from the audio input signal may be obtained. A positive bias is applied to the input signal, V_s, so that the emitter diode polarity is satisfied.

The value of the input resistor R1 should be chosen so as to limit the modulating input current to a maximum of 5 mA. The 100 ohm load resistor of the phototransistor results in rather a low gain, but the 741 stage provides a gain of about ten so that a reasonably large output voltage is obtained.

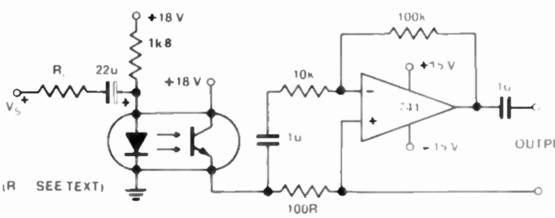


Figure 12. An isolation circuit covering the whole audio range

The low value of the collector load resistor enables an upper frequency up to 20 kHz to be obtained, while the lower frequency response is determined by the values of the coupling capacitors employed — about 25 Hz in the case of the values shown.

Two separate + 18 V supplies are required if complete isolation between the two parts of the circuit is needed. The input resistor R_1 may consist of a variable resistor in series with a fixed resistor if it is required to alter the output signal voltage without any danger of receiving an electrical shock from the output circuit when the latter is at a relatively high voltage.

TTL interface

Optocouplers are widely used in interface logic circuits where the logic signal must be transferred from a circuit at either a high or a low voltage level to a circuit at a very different voltage level.

The circuit of Figure 13 shows how an optocoupling device employing a simple output transistor may be employed to couple the output of a TTL gate to one of the inputs of a TTL 7413 device at a very different voltage level. The 7413 Schmitt circuit provides switching.

A Fairchild report suggests that the base of the output phototransistor of the optocoupling device should be connected to the emitter through a resistor of about 200 kilohm to prevent false triggering of the outputs.

Another logic circuit for coupling an input to a 7413 device is shown in Figure 14, but in this case the 4N33 with its photo-Darlington output device is used.

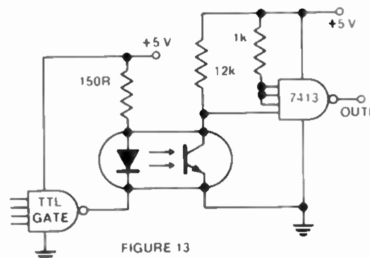


Figure 13. Isolating TTL circuits with an optocoupler.

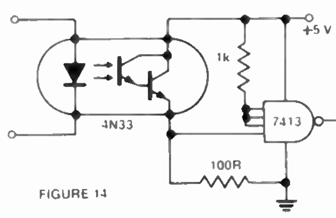


Figure 14. Control of a TTL Schmitt trigger circuit from a 4N33 photo-Darlington device.

It may be noted that in Figure 13 the load resistor (12 kilohm) is much higher than in Figure 14 (100 ohm), but the use of the higher gain of the 4N33 makes up for the lower value of load resistor.

Simple latch

The very simple latching circuit of Figure 15 can employ a pair of 4N33 photo-Darlington output devices. Initially, S1 is open and no current flows through either 4N33. If S1 is then closed, a current flows from the positive supply through the diode emitter in the upper 4N33 and through the emitter in the lower 4N33, the output of the upper device being shorted out by S1 during this time.

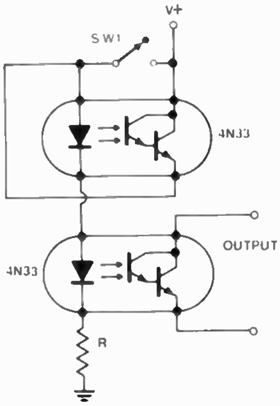


Figure 15. A latching circuit using two 4N33 devices.

When S1 opens, the short is removed from its output circuit, but the response time of the latter is longer than that of the emitter. The current therefore flows through the output of the upper 4N33, through the diode emitter of this same device to maintain the output in its conducting state and through the emitter of the lower 4N33. Thus the output of the lower device remains in its conducting state after S1 has re-opened.

The voltage across the two forward-biased emitting diodes is around 3.5 V and it is convenient to operate these diodes at about 5 mA. Thus, a suitable value for the resistor R is (V+ − 3.5)/0.005 or about 3.9 kilohm with a 24 V supply.

Bidirectional control

The output current of an optocoupler using a phototransistor or a photo-Darlington device must flow only in one direction, so such a device cannot control alternating current.

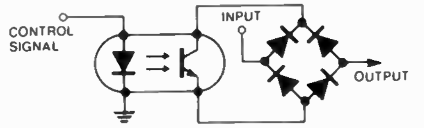


Figure 16. Controlling a bi-directional current using an optocoupler.

This problem can easily be overcome by the use of the circuit of Figure 16, in which the input-to-output current is rectified by a diode bridge circuit before being fed to the output stage of the optocoupled device.

The control signal which switches the output on and off must be unidirectional.

Power supply

Optocoupling devices can be used to isolate the control voltage of a regulated high voltage power supply from this supply line. The basic circuit which may be used is shown in Figure 17.

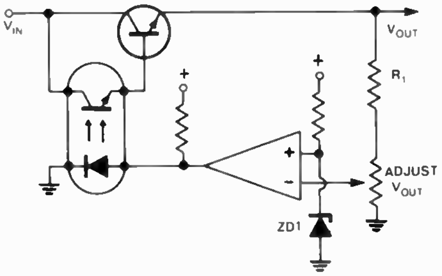


Figure 17. Using an optocoupler in a high voltage series-pass regulator.

A current flows from the stabilised output supply through the high value resistor R1 so that the variable resistor taps off a voltage proportional to the output voltage. This is compared with that across the zener diode D1 using the operational amplifier.

The output signal from this amplifier is fed to the emitter of the optocoupled device which is used to control the series pass transistor and hence to keep the output voltage constant. Thus, the amplifier device output is isolated from the high voltage supply.

A photo-Darlington device may be used in this type of circuit for higher feedback loop gain, but an external pass transistor is always required, since the output devices incorporated into optocouplers can handle only very limited power.

Fast interface

The 6N139 with its 'split photo-Darlington' output device enables the high speed of the separate photodiode to be combined with the high gain of the Darlington connected internal transistors. Although the CTR has a minimum value of 400% at a 500 mA input current, the device output can switch in a few microseconds.

A fast non-inverting logic interface circuit using this device is shown in Figure 18. The maximum switching speed depends on the load resistor, R2, and the input resistor, R1. If R1 has a value of 180 ohm a current of about 17 mA will flow to the output of the TTL input device from the internal emitter diode and the use of a 100 ohm load resistor for R2 will then enable data rates of about 300 kbit/s to be obtained. On the other hand, R1 may be increased to 1k8 for a 1.7 mA diode current with R2 2k2 for a maximum data rate of nearly 50 kHz.

Electrocardiograph amplifier

The use of an optocoupled device to provide complete isolation of a patient from electrocardiography equipment is shown in Figure 19. The electrodes from the patient are connected to the programmable 4250 preamplifier stage which operates from +/−3 V battery supplies, nulling facilities being provided by the variable resistor connected between pins 1 and 5.

The same +3 V battery supply provides the bias for the high gain BC109 transistor which drives the diode emitter of the optocoupling device.

The output phototransistor of the optocoupler receives a base bias so that some current is always passing through its collector circuit. This enables the positive and negative parts of the signal waveform to be obtained at the output.

This is a particularly important application of optocoupled devices, since without the isolation provided by such a device, small currents could be fed into the patient which in certain circumstances could produce death.

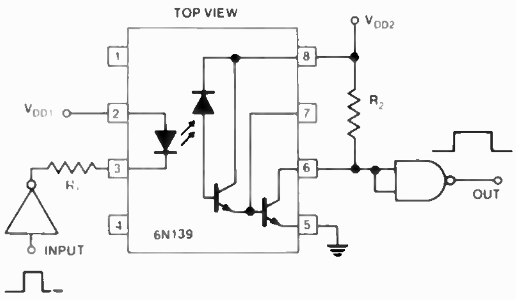


Figure 18. A fast TTL interfacing and isolating circuit using the 6N139.

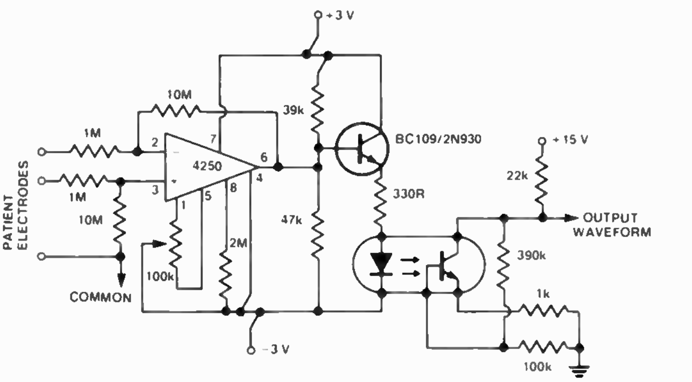


Figure 19. An electrocardiograph preamplifier circuit providing isolation of the patient from the equipment. (Litronix.)

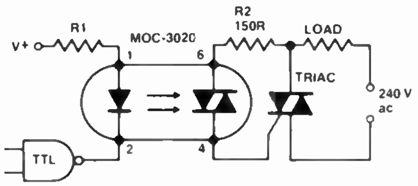


Figure 20. Control of ac power where there is a resistive load, using the MOC-3020.

The MOC-3020

The small triac in the MOC-3020 output can provide a current of up to 100 mA. This is too small for controlling the mains current passing through the load in almost all applications, but is adequate to trigger an additional external triac.

A circuit of this type is shown in Figure 20 in which the output of the TTL gate, controls the emitter current of the MOC-3020 which triggers the internal triac, the latter triggering the external triac.

The latter device should be selected so that it can hold-off the applied mains voltage and also pass whatever current is required by the particular load being used.

Figure 21 shows the use of the MOC-3020 to switch the ac current through a lamp fed from the 240 V mains when the lamp current is less than 100 mA. As the filament of the lamp has a much lower resistance when it is cold, care must be taken to ensure that the initial peak current is not excessive (about 1 A for a very short time is permissible).

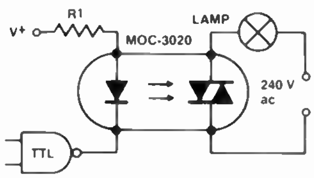


Figure 21. Controlling a lamp on the ac mains using a MOC-3020 (but watch the power rating).

In the circuits of Figures 20 and 21, the load is resistive and conduction of the internal triac ceases when the mains voltage passes through zero during the course of the mains cycle.

In the case of an inductive load (such as an electric motor), however, large back-emf pulses can be generated when the current ceases to flow through the load and this could cause the internal triac of the optocoupler to operate in an improper way.

This problem can be avoided through the use of the type of circuit shown in Figure 22, the values of the components of the 'snubber network' connected across the external triac being dependent on the load inductance and resistance.

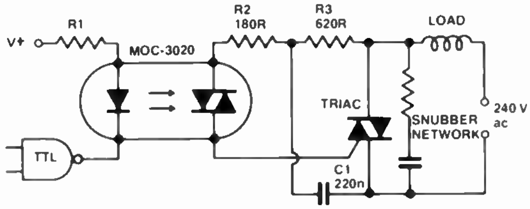


Figure 22. Control of ac power where the load is inductive (i.e: a motor), using the MOC-3020. Note the use of a 'snubber' network. Typical values for the RC network would be R = 180 ohms, C = 220n.

Conclusion

Simple optocoupler devices can be employed in a wide range of circuits from the simplest types to quite complex ones. At prices ranging from under one dollar up to a few dollars, they are excellent value! ●

Measuring receiver performance by the SINAD method

By far the best way to measure receiver performance, especially during alignment, is to employ the "signal plus noise and distortion", or SINAD, technique. This article explains the technique and the pitfalls of 'traditional' methods and introduces a newly available instrument to speed the task.

Peter D. Williams
Vicom International

THERE SEEMS TO BE a general lack of appreciation, even carrying through to some at the engineering levels, about the importance of receiver sensitivity degradation in total communications system performance. A lot of it is caused by the fact that receiver sensitivity is specified in microvolts across a given input resistance, while transmitter power is given in watts.

Few service technicians would let a 30 W transceiver out of the shop that only delivered 15 W output. At the same time, many would not consider the sensitivity unduly low if it measured 0.5 microvolts in a set rated at 0.35 microvolts. Bringing the transmitter back to specifications is likely to require replacement of an expensive power transistor, but bringing the receiver back to specifications will usually require nothing more than a little alignment correction.

The typical land-mobile repairman will always measure transmitter power output before a radio leaves the workshop — and managers insist on pretty good accuracy in their RF power meters. Receiver sensitivity measurement, however, is often merely estimated by ear.

Time and money are part of the reason for this neglect. Sensitivity measurements take time and commercial organisations are facing increasing crises in hiring enough skilled technicians to keep up their business. Even those operated by a government entity, or a large company doing their own communications maintenance — are seeing the pinch. But the number of radios to maintain continues to increase faster than the number of technicians required to support them.

SINAD — the most meaningful

It is pretty well acknowledged that the SINAD method is the most meaningful one for measuring FM receiver performance. It has the basic requirements of any good performance measurement: (1) it is repeatable; (2) it is quite insensitive to variations in technique by those making the measurement; and (3) the measurement relates directly to the actual, practical performance of the receiver.

The SINAD measurement is generally accepted by industry. Sensitivity requirements as stated in Department of Communication specifications gives SINAD as the way of stating useful sensitivity. You seldom see the old 20 dB quieting methods in manufacturer's specifications any more, except as back-up to the universally stated "12 dB SINAD".

Let's review the procedure for making a 20 dB quieting measurement on a receiver. The receiver squelch is set "open", and an ac voltmeter is connected to the speaker terminals to read the level of "thermal" noise delivered by the receiver. The volume control is set to obtain a handy reference level on the ac voltmeter (usually 0 dB). An unmodulated (CW) signal on the channel frequency is then introduced at the receiver input and increased in level until the noise output at the loudspeaker drops to one-tenth (20 dB) of its previous level.

The trouble with this method is that it only measures the ability of the receiver to receive a CW signal. A receiver must receive voice-modulated signals if it is to be useful in land-mobile applications. In order to receive a modulated signal, the receiver must have an adequately flat bandwidth to properly receive the components of the modulated signal.

Poor design, component aging or failure, or improper alignment can result in a peaked response that admits the CW signal nicely but will not permit effective reception of the modulated signals. A signal with a peaked response may seem exceptionally good when measured by the 20 dB quieting method.

The SINAD method, on the other hand, provides an unambiguous measure of the ability of a receiver to receive a modulated signal. Unfortunately, this method has the reputation for being a pretty time-consuming procedure.

The SINAD method

To measure sensitivity by the SINAD method, a signal generator is connected to the receiver antenna terminals and set exactly on the channel frequency. (See Figure 1). The generator signal is modulated by a 1000 Hz tone, and the peak modulation deviation is set at 3 kHz (for receivers used in systems with 5 kHz deviation). The receiver volume control is adjusted to deliver the receiver's rated

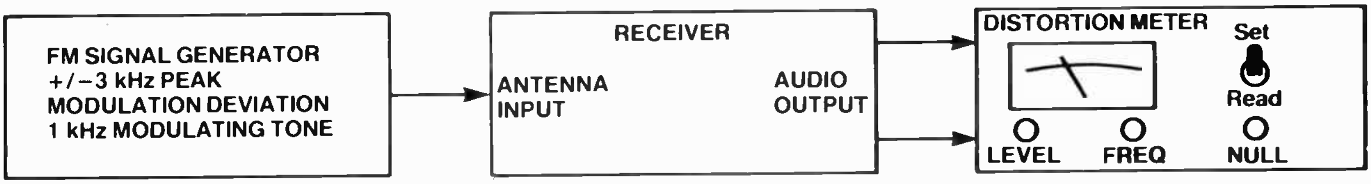
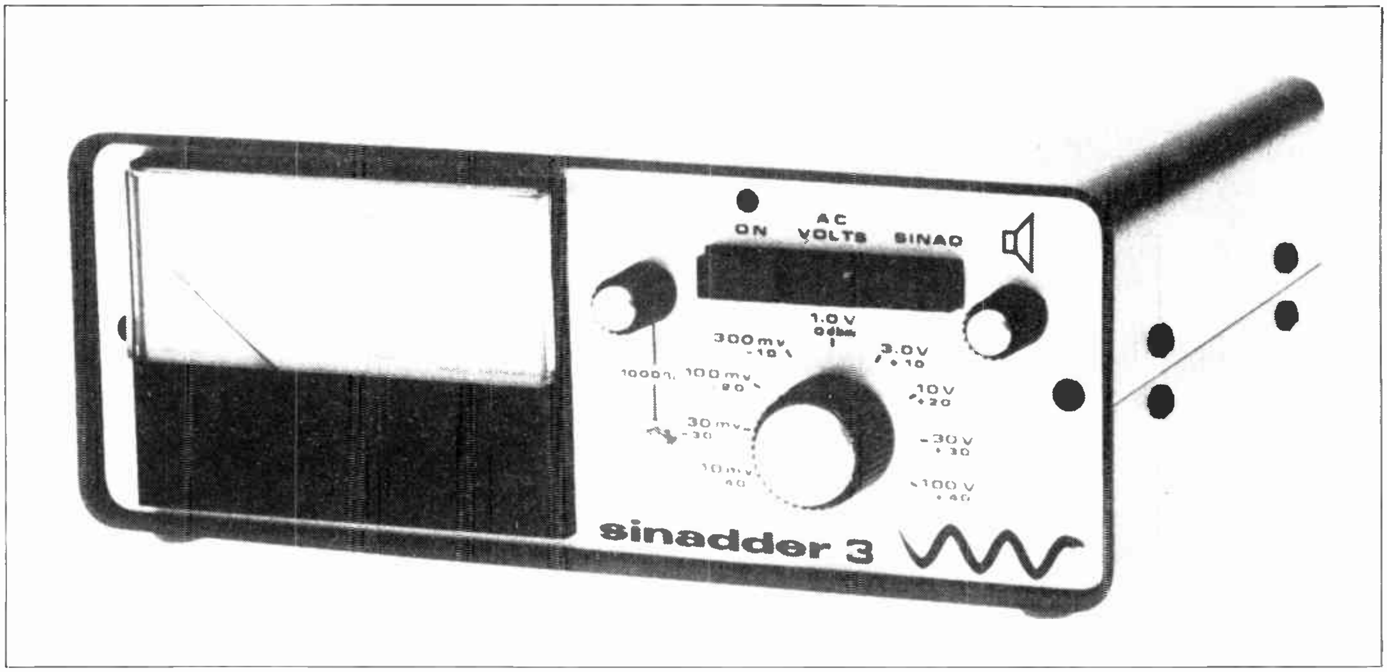


Figure 1. The 'standard' setup for measuring SINAD performance of a receiver.



audio output power, and the distortion meter is connected to the audio output terminals. The distortion meter reference level control is set and the meter is then switched to read distortion. The frequency and null adjustments on the meter are adjusted to null out the 1000 Hz tone. The signal-generator attenuator is then adjusted to obtain a distortion meter reading of 25% (corresponding to 12 dB). The microvolts reading on the signal-generator attenuator is then the "12 dB SINAD sensitivity" of the receiver.

The basis of this procedure is the following: The distortion meter is being used as an audio voltmeter. When the reference level control is being set, the voltmeter is reading all of the components of the audio output of the receiver.

This audio output consists of: **(a)** the 1000 Hz tone (corresponding to the wanted speech intelligence signal to be received in actual operation); **(b)** harmonics of the 1000 Hz tone (distortion); and **(c)** noise — the thermal noise you hear when a weak signal is being received.

When the distortion meter is switched to the "distortion" position, a null circuit filters out the 1000 Hz tone, leaving the distortion products and the noise. The meter is usually calibrated in per cent (%).

When a distortion meter is used as above in a SINAD measurement, it is giving the answer to the following equation

$$\text{Meter reading (\%)} = \frac{\text{(noise and distortion)} \times 100}{\text{(signal + noise and distortion)}}$$

Distortion meters are usually calibrated in per cent distortion, but SINAD measurements are customarily stated in decibels. A 25% reading corresponds to 12 dB, and a 10% reading corresponds to 20 dB. The 12 dB SINAD sensitivity is almost universally used. This 12 dB level is especially appropriate for land-mobile receivers because incoming signals become usefully understandable at levels above about 12 dB SINAD. That is, 12 dB SINAD represents a minimum for an intelligible signal.

As you can see, SINAD measurement can be time-consuming. Most workshops that use SINAD measurement don't worry about adjusting the receiver to its rated audio output. This is a justifiable shortcut, because the audio distortion in any decent audio system will make a minor change in the measurement. Even with this time saving, however, there is still a lot of knob twiddling to do. One problem is that the total output of the receiver may vary somewhat at low signal levels, making it necessary to check or reset the reference level control several times during a measurement.

Knob twiddling removed

An instrument is now available which removes the knob twiddling and has a meter reading directly in SINAD. A very stable active notch filter preset at 1 kHz (by EIA definition, the SINAD test frequency) is internally provided to eliminate frequency and null controls. A tight AGC circuit permits the instrument to operate over an input voltage range of from 30 millivolts RMS to over 4 V RMS while maintaining a constant reference level, thus eliminating the level set control and reference-distortion switch. This leaves an on-off switch as the only control on the panel. The unit draws less than 5 W and can be left on indefinitely. To measure SINAD with this instrument, all you have to do is connect it to the receiver loudspeaker leads and feed a measured, 1 kHz modulated signal into the receiver front end.

The 'SINADDER', as it is called, is made by the Helper Instrument Co. and proves to be just as much of a time saver as expected. Figure 2 shows the procedure for measuring 12 dB SINAD sensitivity using the instrument. Assuming a 3.2 ohm loudspeaker, the meter will read correctly from a receiver output level of a quarter of a milliwatt to over five watts. That is plenty of range when you consider that the lower level would be hard to hear and the higher level would probably drive everybody out!

As an alignment aid

Although the instrument was designed to measure SINAD sensitivity, it is also a tremendous alignment aid. It is this use that should make it popular in a lot of service shops. Typical alignment procedures for FM receivers consist of dc metering at specified meter points. Usually a weak signal is injected and the alignment adjustments are made to maximize the meter readings. Alignment "by-the-meter" as most technicians know, is not always the optimum alignment.

These by-the-meter procedures give the alignment for maximum gain of the various stages, but this is not necessarily the alignment for optimum signal-to-noise performance. It is often possible to improve on a by-the-meter alignment by touching up a few of the adjustments in the front end for optimum signal-to-noise.

Although a few of the old hands can get good results doing this touchup by ear, the SINADDER makes the touch-up procedure really practical — and fast! By retuning for optimum SINAD, it is almost always possible to squeeze a decibel or so extra sensitivity out of a receiver, and a 3-to-5 dB improvement is not at all unusual.

The automatic reference level control circuits in the SINADDER make the "alignment for best SINAD" procedure possible. After the receiver is roughly aligned by the usual methods, the signal generator is set to provide a 1 kHz ▶

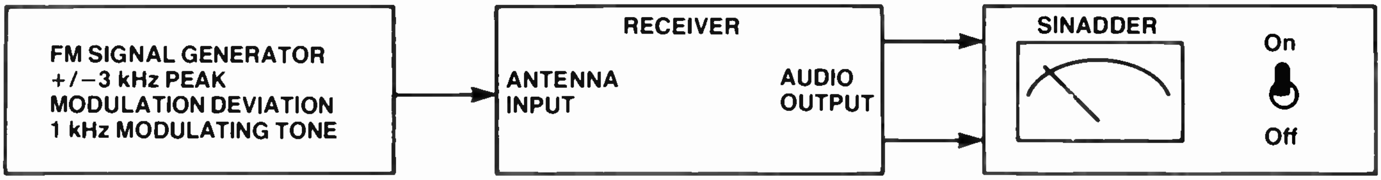


Figure 2. Using the 'Sinadder 3'. Note that there are no adjustments to make on the Sinadder, hence 'knob twiddling' is removed.

modulated signal of the correct deviation and the alignment adjustments are touched up to obtain minimum deflection on the SINADDER meter.

As the procedure progresses, the signal generator is backed off to keep the SINAD reading at about 12 dB. This procedure avoids the peaked response that often occurs in the ordinary meter-tuning methods. It results in a better bandpass alignment of overcoupled circuits and crystal filters than is obtained with the ordinary limiter-meter procedures.

"Needle nose" aligning

One of the recurring problems in connection with receiver alignment concerns the occasional receiver that ends up with a "needle nose" bandpass. This can be caused by faulty receiver design, or by aging of bypass components, or a host of other reasons. Although it would be best to get truly to the bottom of the problem, the pressure of time often makes it necessary to get on the next job, and it is usually possible to arrive at an alignment adjustment that results in normal performance.

If the alignment of one of those "needle nose" jobs is touched up for optimum SINAD, normal performance can often be achieved. This procedure is aided if the modulation deviation on the signal generator is set at about 5 kHz instead of the 3 kHz that would be used for a SINAD measurement.

Modern receivers are making increasing use of quartz crystal filters. Sometimes these have tuning or matching adjustments associated with them. In general, they cannot be tuned by the usual adjustment for maximum limiter current. Some manufacturers specify a procedure that requires a sweep generator and a 'scope'. It takes quite a bit of time just to set up everything for the sweep alignment. It is possible to arrive at proper alignment of these crystal filters with the SINADDER.

Set the signal generator for 5 kHz modulation deviation, reduce the signal generator output until the SINAD indication is about 12 dB, then adjust the tuning controls for minimum deflection on the SINADDER meter. The resulting alignment is superior to the one obtained by the sweep method. This is because the sweep method arrives at the adjustment for a flat amplitude characteristic in the passband, whereas the SINADDER adjustment leads one to the most linear phase characteristic — which is more important to the FM detector.

Frequently, a technician is confronted with a receiver that needs alignment and the radio is a model with which he is unfamiliar, and the instruction manual is not on hand. Alignment then becomes a pretty tricky proposition. Using the conventional limiter-metering approach, it's easy to go astray because you are never sure whether you are twisting an adjustment that is located after the test point you are observing, or whether some intervening limiter is masking the effect of the tuning.

The whole thing is a lot less tricky when you align the receiver for optimum SINAD. You know where the antenna input and the loudspeaker terminals are, so you can't make the metering mistake mentioned above. If you are trying to do one of these "blindfold" alignment jobs, you should proceed carefully and not make any radical changes in any of the adjustments. The main pitfall to avoid is one of those oscillator circuits which tune with a "cliff" on one side, and won't start when peaked up to the maximum. Otherwise you can usually obtain a pretty good alignment without the book.

AM and SINAD

One final note: if you have been thinking about using the SINAD method of sensitivity measurement for those AM receivers in your shop, you're right; it is just as meaningful as with FM. The 10 dB SINAD sensitivity of a decent receiver is surprisingly close to the 10 dB signal-plus-noise-to-noise specification often seen with AM receivers. You can also use the "alignment by SINAD" method to get the last bit of sensitivity out of the AM receiver. Any touch-up of the IF tuning probably won't gain much, but the front-end touch-up will usually show a worthwhile improvement. ●

THE SINADER 3 — A BRIEF REVIEW

Roger Harrison

I couldn't resist the opportunity to get hold of a Sinadder and try it for myself. Having done quite a few SINAD receiver measurements over the years with the 'usual' collection of gear — RF signal generator, audio oscillator and noise-and-distortion meter, I just had to see if this unit would deliver what it promised.

The Sinadder 3 comprises a 1 kHz precision oscillator, an ac voltmeter and a SINAD measuring circuit. The latter consists of a gain-controlled amplifier, which maintains a constant average output level for inputs ranging between 10 mV and 10 V RMS, followed by a 1 kHz notch filter and a precision rectifier driving a calibrated meter. A loudspeaker amp. stage is also included.

The ac voltmeter section consists of an input attenuator followed by a x100 amplifier stage driving the precision rectifier and calibrated meter.

The 1 kHz oscillator employs a low-distortion three-stage phase-shift circuit followed by an amplifier/buffer stage.

A voltage-regulated mains power supply is provided but the unit can be powered from an external 12 V(nom) source.

Three interlocked pushbutton switches on the front panel select the function. The 1 kHz oscillator output is brought out to a BNC socket on the front panel. Both the oscillator output level and the internal speaker level can be adjusted by front panel controls.

The input lead comes through the rear apron, via a clamp grommet, and consists of a shielded cable about one meter long with alligator clips on the end. The SINAD input level can be between 20 mV and 10 V RMS for correct operation; input impedance is given as 100k and accuracy in SINAD mode is quoted as ± 1 dB. The ac voltmeter has nine ranges from 10 mV to 100 V in 10-30-100-300 etc steps. The input impedance is given as 1M and accuracy quoted as ± 3% of full scale, ± 0.25 dB, 100 Hz to 20 kHz.

The 1 kHz oscillator is quoted as being within ± 1 Hz, output 1.5 V RMS into a 500 ohm load.

The unit measures 222 mm wide by 82.5 mm high by 178 mm deep. It is manufactured by the Helper Instruments Co of Florida, USA, and distributed in Australia by Vicom International, 57 City Rd, South Melbourne Vic. (03) 62-6931. Recommended retail price is $395, plus tax.

The instruction manual supplied with the unit is comprehensive and covers use of the instrument in detail, together with some background on SINAD measurements. A complete circuit and board overlay of the instrument is included along with a description of its operation. Warranty is for 12 months.

The Sinadder 3 is just so damned easy to use compared to what I've been used to. As a trial, I set up a new UHF CB rig we have in for review and measured the receiver SINAD sensitivity figure using our Hewlett-Packard 8654B RF signal generator and AWA F242A N&D meter. It took me 22 minutes (I'm out of practise!).

I then switched to the Sinadder 3, hooking its 1 kHz internal oscillator into the H-P generator's external FM input. The two readings were within 0.01 uV of each other. The receiver measured 0.26 uV SINAD (a pretty commendable performance). With the Sinadder 3, it took me just 10 minutes (no practise!).

The Sinadder 3 represents excellent value for money and would be a useful tool in any RF experimenter's workshop, communications service workshop or even a communications field serviceman's toolkit. Recommended.

THD analyser for audio circuits

This article describes a spot frequency audio distortion analyser, designed and built by ETI reader, Laurie Tunnicliffe of Mulgrave, Victoria. Measurements can be made at 100 Hz, 1 kHz, and 10 kHz. The final resolution of the instrument is 0.01%.

IN RECENT YEARS there has been a trend towards considering the transient behaviour of audio circuits rather than their steady-state behaviour. Although the attention given to this side of circuit design is not unwelcome, total harmonic distortion (THD) analysis is far from redundant.

While the transient behaviour is a go/no go situation, the THD behaviour is a measure of how well a circuit will perform. For instance, it is not a question of to what degree an amplifier will slew limit, or to what extent the internal loop will overload; these transient characteristics are barriers, and until reached will have no effect on the amplifier's performance.

THD measurements are therefore still a valuable analytical tool when developing new circuits or measuring and giving figures of merit to existing circuits.

When a single frequency is passed through an amplifier with a non-linear transfer curve, other frequencies are produced. These other frequencies are integer multiples of the test signal and sum of these is the THD.

There are a number of ways of measuring the harmonic distortion of an amplifier. The most recent is the 'Fourier Analyser', which is a computer-based instrument that samples the output waveform and performs the necessary mathematics (Fourier transform) to break down the waveform into its component parts. These instruments however cost tens of thousands of dollars. A technique becoming popular is the use of a spectrum analyser, which is a swept bandpass filter. The results are displayed on a CRO. THD is then calculated from

$$\text{THD} = \sqrt{\left(\frac{F1}{F}\right)^2 + \left(\frac{F2}{F}\right)^2 + \left(\frac{F3}{F}\right)^2 \ldots}$$

N.B. Square root sign in formula applies to sum of all terms.

Another technique used by Quad and described elsewhere[1], compares the input and output of an amplifier (using a differential amp), the difference being distortion. In Quad's experiment, music is used as the source and a monitor amp/speakers are used to listen to the distortion played by itself. This then becomes a 'real-time' distortion analyser, and any non-linearities, whether transient or steady state, are revealed.

The last and most commonly used method is to eliminate the fundamental signal and read the resultant harmonics on a moving coil meter. This is sometimes called 'Noise and THD' measurement, as hum and 'electronic noise' are lumped together with the harmonies. This is the technique used for the instrument presented here.

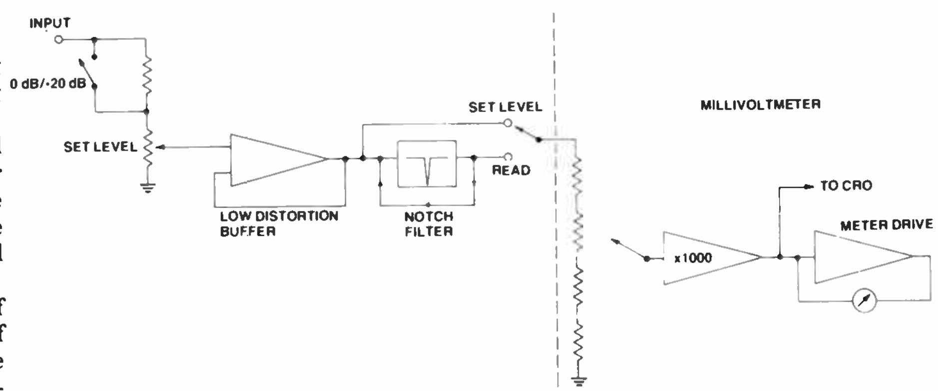


Figure 1. Block diagram of the THD analyser described in this article.

Block diagram

Refer to Figure 1 for the instrument's block diagram.

The input is applied to a buffer stage via a 0 dB/−20 dB attenuator. This allows easier control of the 'set level pot' when large signal levels are being measured. The buffer stage provides a low impedance source to drive the notch filter. The design of the buffer is of major consideration, as any non-linearities it contributes will be indistinguishable from the signal being tested.

There are a number of options available when designing a notch filter. A derivative of the Wien bridge was chosen due to its simplicity and the fact that it only requires two variable reactances to balance.

Bootstrapping around the filter is necessary to tighten up the notch width. Without this, the attenuation of the second harmonic would be excessive. With the amount of bootstrapping used the second harmonic is attenuated by less than 1 dB.

The notch filter is followed by a millivoltmeter and reads the average value of the harmonics relative to the fundamental. The meter is calibrated to read full scale for 0.775 V RMS input, and therefore the meter can be used separately to measure dBs into a 600 ohm load, relative dB (dBV) or millivolts, giving the instrument a dual function.

A CRO output is taken after the x1000 amplifier, so that the residual harmonics can be investigated. This will often give considerable insight into the cause of the distortion. ▶

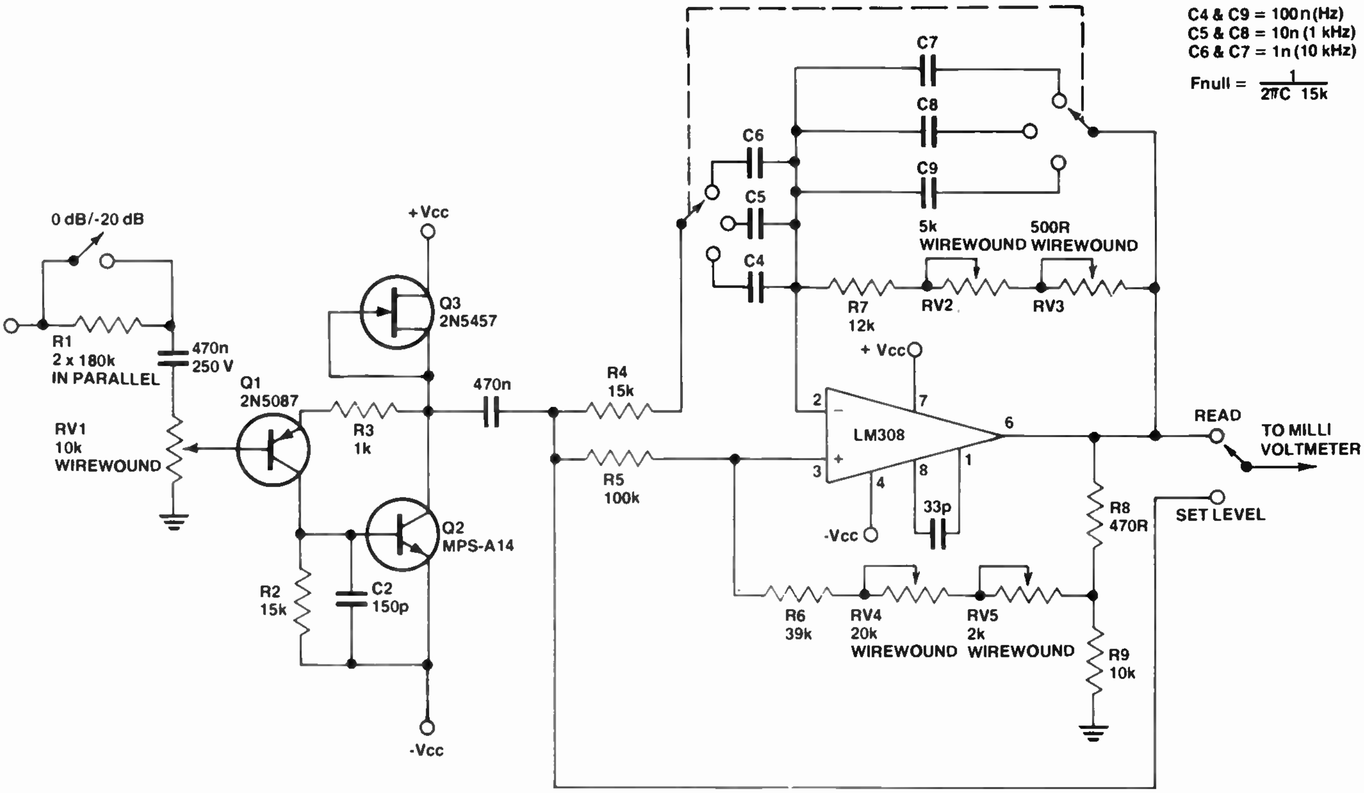


Figure 2. **Circuit diagram of the THD analyser (above and opposite).**

Circuit description

By examining Figure 2 the complete circuit diagram can be understood.

Q1 and Q2 form the non-inverting buffer with Q3 acting as an active load for Q2. The distortion contributed by this stage can be calculated to a first approximation as follows. The input transistor contributes negligible distortion relative to the second stage, due to the small signal levels it handles. The second stage is the prime mover, with a voltage gain of approximately 60 dB. The base drive voltage will therefore be

$$\frac{.775 \times \sqrt{2}}{1000} = 1.1\,\text{mV}$$

peak and the distortion generated is 1.1% second harmonic[2]. Since the buffer stage has 100% feedback (unity gain), the loop gain is also 60 dB and the distortion is reduced to

$$\text{loop gain} = \frac{1.1}{1000} = .0011\%$$

The buffer feeds the notch filter, which may be looked upon as a frequency dependent differential amp. At the notch frequency, the parallel arm and the series arm balance to give the same impedance ratio as the resistive arms. The input then appears as a common mode signal to a differential amp, and the output is zero. The common mode rejection ratio of the op-amp is of particular importance; however, most op-amps have a CMRR of at least 80 dB, which is sufficient to give a 0.01% resolution.

R8 and R9 provide the positive feedback (bootstrap) necessary to tighten the notch width.

Following the notch filter is the millivoltmeter, which consists of a constant gain, x1000 amp, preceded by a step attenuator giving 20 dB steps. The meter ranges will therefore be 100%, 10%, 1%, 0.1%.

The amplifier is followed by a fairly conventional meter-driven circuit, with the four bridge diodes being placed in the feedback loop of the op-amp. The non-linearities of the diodes are rendered insignificant and the meter reads accurately, even at the low end of the scale. Diodes 5 and 6 are used to protect the meter from an overload.

In order to achieve a wide enough bandwidth for the x1000 amp, an externally compensated op-amp was necessary. The op-amp used in the meter circuit is a dual low-noise device, and therefore helps to keep the instrument noise level low and reduces parts count.

C12 rolls off the frequency response above 70 kHz. This helps improve the square wave response by reducing any ringing, and also reduces high frequency noise. This will allow measurement up to the seventh harmonic of 10 kHz, and should prove adequate.

Construction

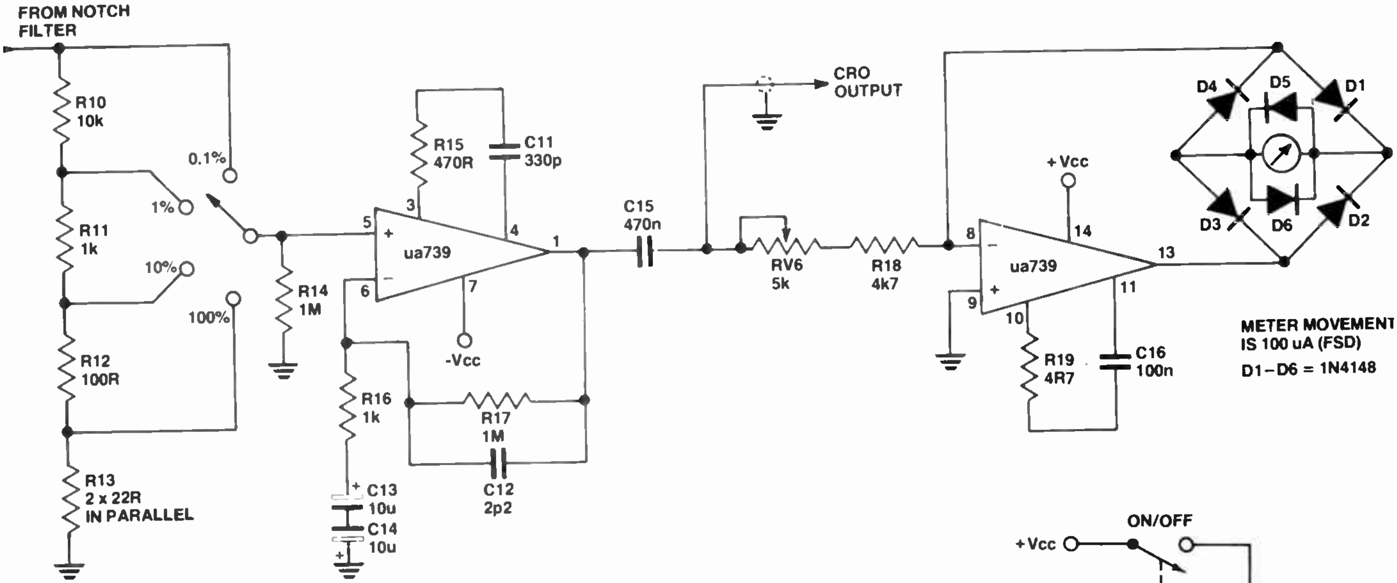
If accuracy and stability are to be reproduced, all components must be close tolerance, high stability. I used ¼-watt metal film resistors, which are available from Dick Smith. The filter capacitors are either styroseal or green caps; ceramic capacitors should be avoided as they are voltage dependent.

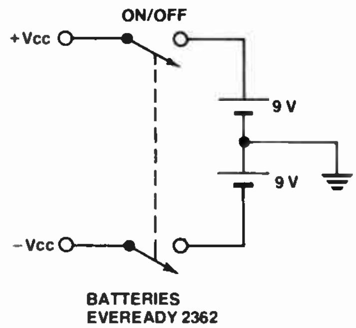
Veroboard should also be avoided, as stray capacitance across the strips causes problems when you are looking for one part in 10 000.

I used tagstrip and wire-wrap sockets for the ICs and found this easy to work with, giving satisfactory results.

My prototype was built in a diecast box; however, I had to rewire it three times, using shielded cable for every connecting wire, before a stable layout was found. For this reason I strongly suggest building the instrument as two separate units — a notch filter and a millivoltmeter. This will also give the versatility of being able to use one or the other independently.

The nulling pots are wirewound, having the advantage of a continuous track. The continuity of a wirewound pot is limited by the fact that a step in resistance equivalent to one wire winding is the smallest change possible. Carbon pots are certainly worse, as they sometimes jump resistance value mid-track. The fine control pots are single turn and provide reasonable ease of nulling at low levels. However, if you





feel it is worth the extra cost, ten-turn (spiral wound) pots will alleviate having to be careful when adjusting the controls.

The only adjustment to be done is the calibration of the meter. This is accomplished by applying a 0.775 V RMS, 1 kHz signal to the input with the unit switched to 'set level', and adjusting RV6 for full-scale deflection of the meter. Alternatively RV6 and R18 can be changed to 1k and 6k8 (fixed values) with less than 1% error in meter reading.

Operation

The instrument is operated as follows:

- Set Level/Read switch to Set Level
- Range switch to 100%
- Adjust Set Level pot for FSD of meter
- Set Level/Read switch to Read
- Adjust the nulling pots for minimum meter reading
- THD is now read from the meter and range switch.

The nulling procedure is accomplished by starting with the coarse pots, and alternately adjusting them for minimum meter deflection until it becomes difficult to proceed. The process is then repeated with the fine pots. If the user has a CRO available, this will assist the nulling procedure.

Figures 3, 4 and 5 show some oscilloscope photographs of the input and output of the meter. Figure 3 is an under-biased class B amplifier and shows spikes in the residual. This is a common waveform from class B amplifiers and the meter can be used to set the bias level for optimum.

Figure 4 shows second harmonic distortion from a voltage-driven common emitter amplifier. The input signal was 1.1 mV peak and the meter reading was 1.1%. This confirms the analysis used for the buffer stage.

Figure 5 is third harmonic caused by a thermistor-stabilised Wien bridge oscillator. THD was 0.02% at 100 Hz. (This is one of the problems encountered when using a thermistor at low oscillator frequencies.)

It should be noted that a low distortion oscillator will need to be used when making measurements. The residual distortion of the oscillator may set the lower limit to the measurements.

Some performance measurements

Final resolution of the meter was 0.005% at 100 Hz and 1 kHz and 0.01% at 10 kHz. Below this, drift in components' values with temperature, circuit noise, distortion introduced by the buffer and the filter stage and CMRR limitations of the filter all take their toll. However, distortion values of less that 0.01% are purely academic in my view, regardless of what hi-fi manufacturers' sales departments would have us believe.

References
1. P.J. Baxandall, Wireless World, November 1977, pp. 63-66.
2. E.F. Taylor, Wireless World, August 1977, pp. 28-32.

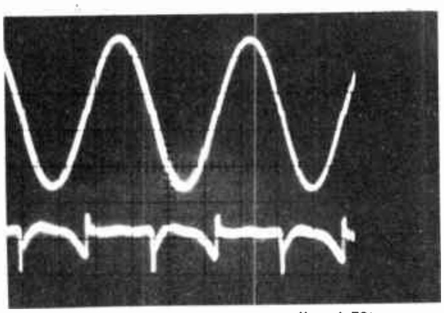


Figure 3. Class B amp, meter reading 1.5%.

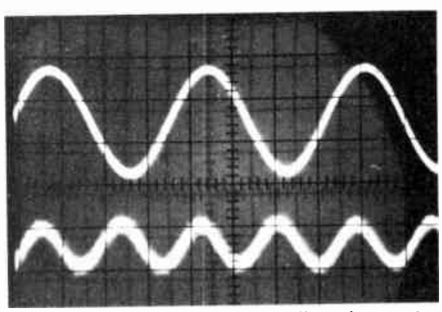


Figure 4. Second harmonic distortion, meter reading 1.1%.

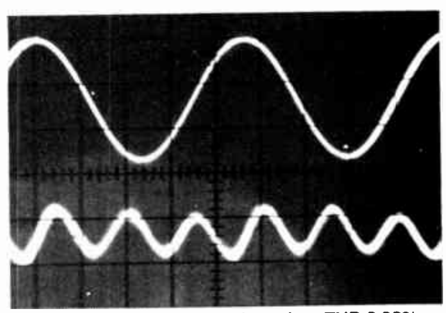


Figure 5. Third harmonic distortion, THD 0.02%

Improved text display for video monitors

Modify your video monitor to uncramp the lines of text. Here are the principles and practical circuitry.

Graham Wideman

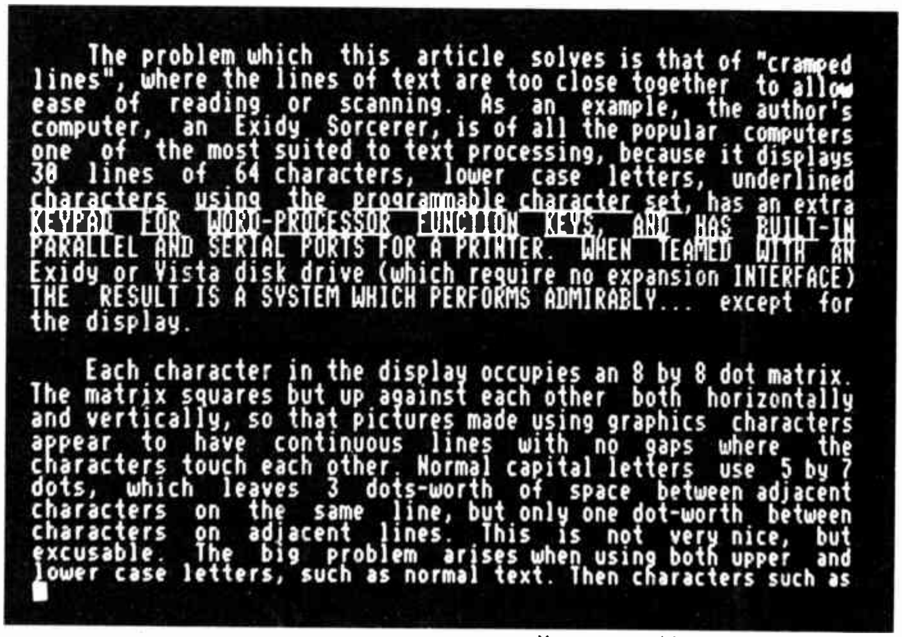
REMEMBERING WHAT we were all playing with ten years ago, there is no denying that today's personal computers are marvellous pieces of equipment, and whatever your brand preference, tremendous value for money. And that sentiment is doubled for those of us who use a computer for 'serious' purposes — in our work. However, the serious user will notice that most personal computers sacrifice some degree of convenience of use for particular specialised applications. This is not surprising, as the manufacturers have tried to make their products as generally appealing as possible without unnecessarily increasing their prices with luxury features.

One such area is that of text display, where many personal computers 'trade in' some text readability in order to accommodate graphics capabilities. For many users this is of little consequence, but it becomes a major headache when using the computer for word processing, or even for long hours of programming. This article describes a method whereby a video monitor can be modified to counteract one of these shortcomings, and provide a display which will save your eyes and your temper.

Cramped lines

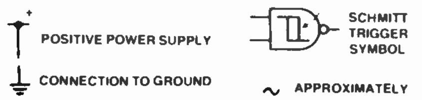
The problem which this article solves is that of 'cramped lines', where the lines of text are too close together to allow ease of reading or scanning. As an example, the author's computer, an Exidy Sorcerer, is of all the popular computers one of the most suited to text processing because it displays 30 lines of 64 characters, lower case letters, underlined characters using the programmable character set, has an extra keypad for word processor function keys, and has built-in parallel and serial ports for a printer. When teamed with an Exidy or Vista disk drive (which require no expansion interface) the result is a system which performs admirably . . . except for the display.

Each character in the display occupies an 8-by-8 dot matrix. The matrix squares butt up against each other both horizontally and vertically, so that pictures made using graphics characters appear to have continuous lines with no gaps where the characters touch each other. Normal capital letters use 5-by-7 dots, which leaves three dots' worth of space between adjacent characters on the same line, but only one dots' worth between characters on adjacent lines. This is not very nice, but excusable. The big problem arises when using both upper and lower case letters, such as normal text. Then characters such as 'y' and 'p', which have 'tails' (descenders) hanging below the line, use the bottom row of dots in the matrix for these tails. Now there is no space between lines, and for example a 'g'



You can turn this screen display into . . .

NOTES AND KEY FOR DIAGRAMS



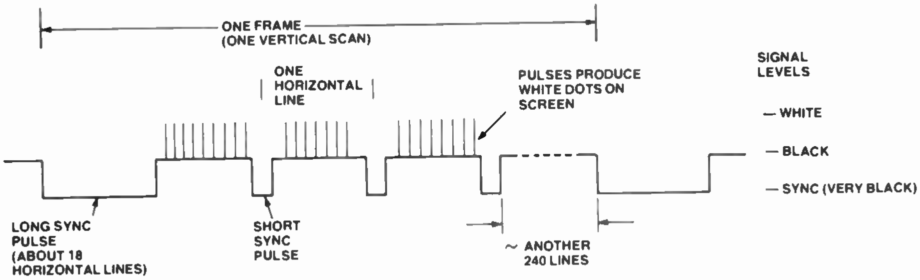


Figure 1: **Simplified diagram of the composite video signal.**

appearing above a 'T' will touch. In addition, underlined characters are most unreadable, since they too use this bottom row of dots in the matrix. It is thus desirable to be able to introduce some more space between each row of characters.

More space

There are several methods which are potentially useful in obtaining the desired spacing. One possibility is to modify the computer so that it allows a picture scan-line (or two) in between each row of characters. This proves to be unfeasible in the case of the Sorcerer because its 30 lines of characters already occupy 240 of the approximately 256 scan lines that you might be able to squeeze onto your monitor, which doesn't provide even one spare line per row of characters. In any case, such a modification to the computer could be rather messy.

Consequently, it is logical next to look at the video monitor itself to see if it could be modified to provide space between rows of characters. The answer is *yes*. The following description of how this is done is *not* a complete design for any particular monitor, but is instead intended to be adapted to whatever monitor you have. The simple circuitry shown was incorporated into a 'monitorised' TV set, and demonstrates the techniques involved.

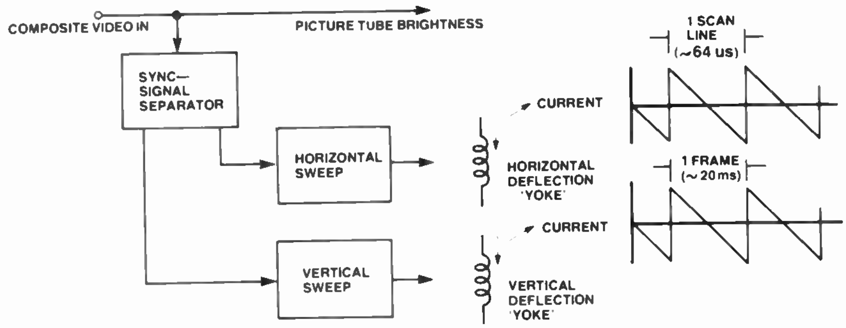
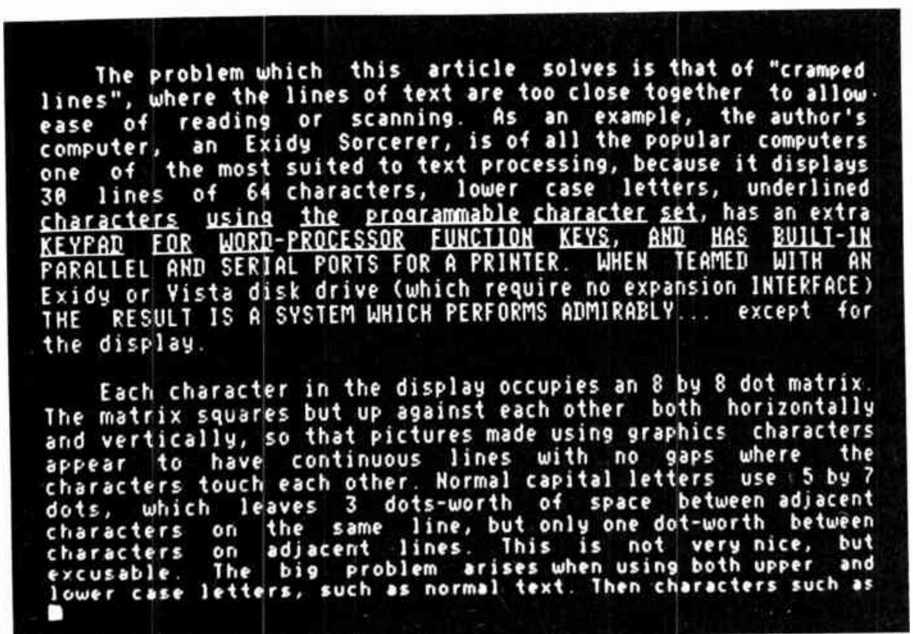


Figure 2: **Conceptual picture of TV or monitor sweep sections.**

Vertical sweep operation

Before proceeding to the design of the 'line uncramper' itself, it is necessary to describe the salient features of the vertical and horizontal scanning systems in a typical monitor. The signal which is passed by your computer to the monitor is called 'composite video' and looks something like Figure 1 when displayed on an oscilloscope. The signal is 'composed' of three separate sets of information added together. They are: the actual picture information which tells where white dots are to appear on a line; the horizontal synchronising pulses which identify the beginning of a line and tell the monitor to start a new line at the left side of the screen; and the vertical synchronising pulses which tell the monitor to start a new frame at the top of the screen. What happens to these signals is shown in Figure 2.

The electron beam inside the picture tube is scanned from side to side and from top to bottom by two sets of electromagnets, called the horizontal and vertical 'yokes' respectively. An electromagnet deflects the electron beam away from the centre of the screen by a distance which is more or less proportional to the amount of current flowing through the magnet coil. Thus it is not surprising that in order to provide a smooth sweep rate over the screen followed by a rapid retrace to the starting position, the current through the coils in both horizontal and vertical cases is a sawtooth wave shape, as Figure 2 shows. ▶



... this screen display. Letters on adjacent rows no longer touch, and a screenful of text (only a small area is shown here) becomes far more easily readable.

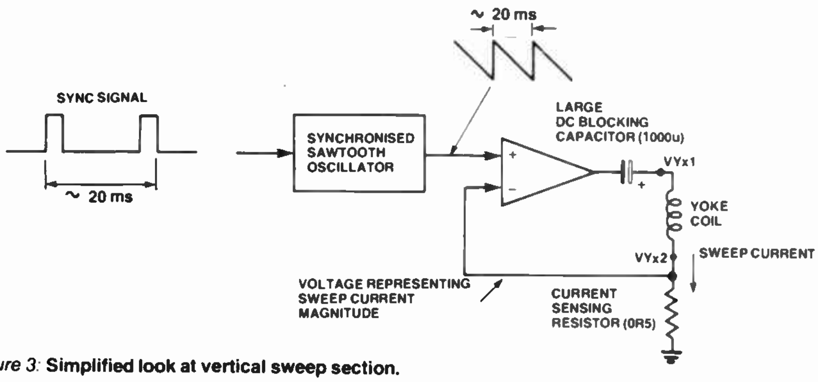


Figure 3: **Simplified look at vertical sweep section.**

A more detailed look at the vertical sweep section is provided by Figure 3. The synchronising signal controls the frequency of the sawtooth oscillator. The sawtooth voltage is amplified, then applied across the vertical yoke coil. The output from the amplifier is a sawtooth-like waveform which varies between, say, 1 V and 7 V. However, the yoke must conduct both positive and negative currents in order to deflect the electron beam up and down to cover top and bottom of the screen. The vertical amplifier cannot generate a waveform with both positive and negative voltages, since it is supplied with only a 'single-sided' positive power supply. Hence a large dc blocking capacitor is inserted so that the voltage applied to the yoke averages out to zero and thus must have both positive and negative parts, say −3 V to +3 V.

A small-value resistor (typically half an ohm) does little to reduce the sweep current through the yoke to ground. Instead, the small voltage generated across it is used as a measure of the yoke current, and is fed back to the negative input of the sweep amplifier. Here it is compared to the intended sawtooth, and any correction is made. The result is that a sawtooth *current* is supplied through the yoke, and a sawtooth *voltage* may be observed at the 'top' of the current-sensing resistor. However, the voltage waveform observed at the top of the yoke (output from the capacitor) is a somewhat distorted sawtooth because the yoke coil has inductance.

This is a simplified description of how the vertical sweep section works, but it points out the three components (yoke, dc blocking capacitor and current-sensing resistor) which are important to the line uncramper. If you are contemplating modifying your monitor then you must be able to locate these three components both inside the monitor and on the monitor circuit diagram.

Sweep modifying

Having recognised that the deflection of the electron beam (and hence the spacing of the scan lines on the screen) is controlled by the current flowing through the yoke, it is straightforward to conclude that we could add a little bit of extra space between the bottom of one row of characters and the top of the next row by somehow altering the yoke current at the appropriate instant.

Figure 4a shows a graph of normal vertical sweep current during a portion of a frame (about 20 lines' worth), and the relative spacing of the picture lines. It would be possible to add current in steps after each character row, as shown in Figure 4b (one character row here is shown as consisting of eight lines, as would be the case for the Sorcerer). This would provide the desired space between character rows; however, all the additional spaces would add to the total picture height, possibly making the picture too high for the screen even with the monitor's picture height adjustment set to minimum. In any case this arrangement would require readjustment of picture height when switching between graphics and text modes.

The basic solution lies in Figure 4c. Here, each set of eight lines belonging to one row of characters is vertically squeezed together. This makes the characters less tall, the skrinkage leaving the desired space between character rows. Figure 4c also shows the waveform of the basic current to be added to the normal yoke current to achieve this result.

The circuitry

Now the actual line uncramper circuitry can be discussed. The circuitry has been broken into sections so that the function of each can be clearly understood, and so that, where necessary, modifications can be made simply. Figures 5 to 11 show these circuit

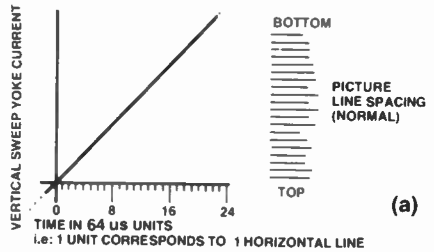


Figure 4a: **Graph showing the vertical yoke current during a portion of the sweep. Beside the graph is a representation of the spacing of the scan lines with this normal (unmodified) sweep current.**

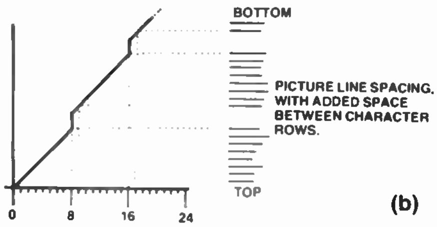


Figure 4b: **Graph shows the effect of adding a step of current after each character row. Note the spaces in the scan lines.**

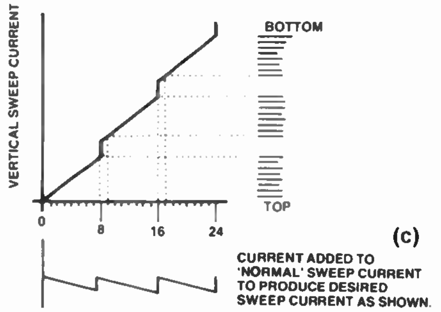


Figure 4c: **In this graph the scan lines for each character row are vertically squeezed together to obtain a space between rows. The current to be added to the normal yoke current is shown beneath.**

modules.

The line uncramper needs to be supplied with power. This can most readily be obtained from the monitor itself, as shown by Figure 5. Needless to say, it is assumed that the monitor or TV set in question is a transistor or IC type and has such voltages available. (Modification of valve type monitors is not recommended, both from safety and practicality viewpoints.) The 'power supply' circuitry filters the supplied voltage so that any noise on the monitor's supply line will not interfere with the line uncramper, and perhaps more importantly, vice-versa. The voltage supplied to the line uncramper can be anywhere between 9 V and 15 V; the components are not very fussy.

Horizontal and vertical signals

The line uncramper needs to be informed of the start of each horizontal line (so it can count them) and the start

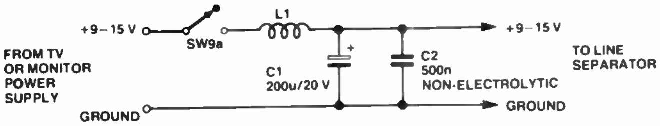


Figure 5: **Schematic of 'power supply' for Line Uncramper.**

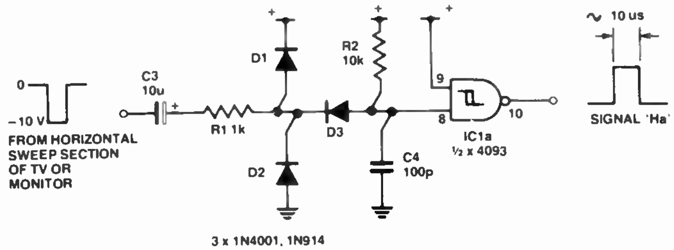


Figure 6: **This circuit is the horizontal pulse conditioner, taking a signal from the monitor and producing a well-defined pulse at the beginning of a scan line for the subsequent circuitry to use.**

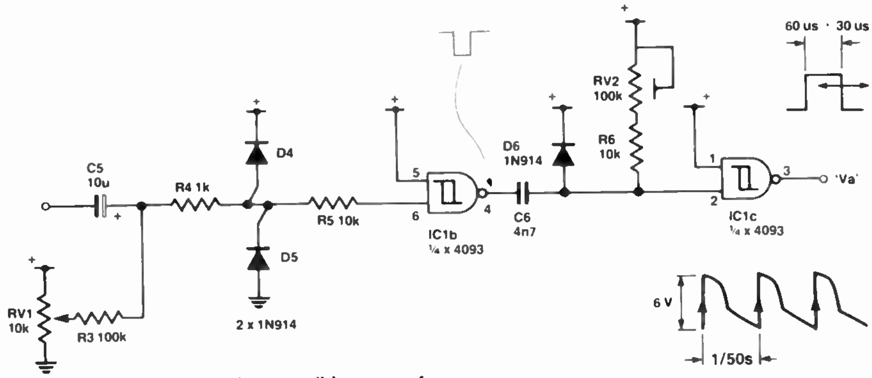
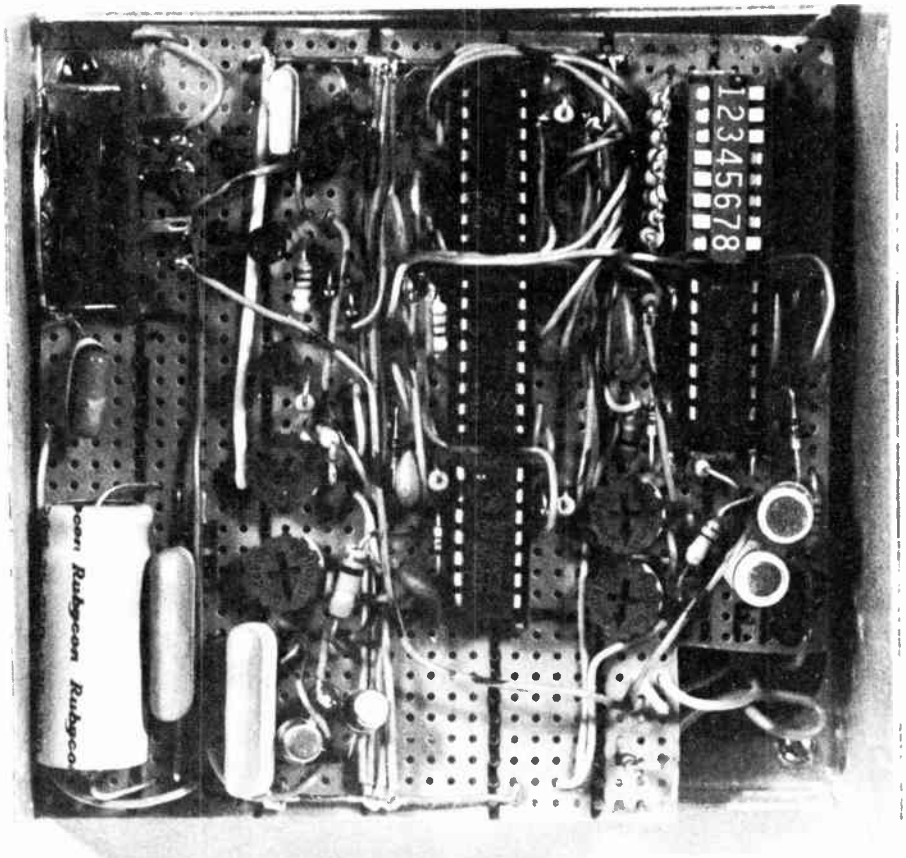


Figure 7: **This vertical pulse conditioner performs an analogous function on the vertical signal to the circuitry in Figure 6.**

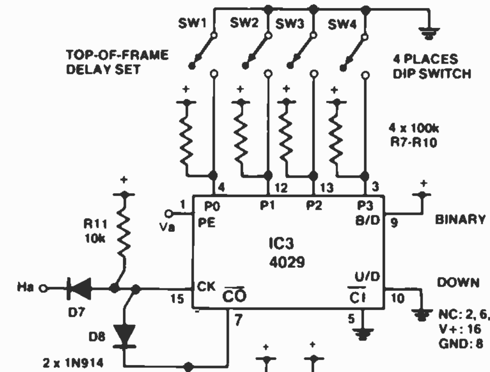
of each picture frame (so it can properly register the spaces between character rows). To facilitate this, signals are taken from the horizontal and vertical sections and conditioned into nice pulses for use by the rest of the circuitry. This job is done by the circuits shown in Figures 6 and 7. In the TV set which the author modified, a 0 V to –10 V pulse was found in the horizontal sweep section. The horizontal pulse conditioner converts this first to a +10 V-to-0 V pulse (components C3, R1, D1, D2) then generates a 10 us pulse on the falling edge of the input pulse (D3, R2, C4) which is inverted and squared up by IC1a. (Note the use of Schmitt trigger-type gates. These give more 'positive' switching action, which is important when attempting to change relatively slowly varying signals into digital pulses. Although the 4093 is shown, any CMOS Schmitt inverters will do).

The vertical signal used was taken from the yoke (point VYx1 on Figure 3). This signal is less 'nice' than the horizontal signal used, and its pulse conditioner consequently is fancier. The waveform is a sort of sawtooth with a pulse on top of each peak, and it measures 1 V to 7 V. This is 'floated' by C5, and the average voltage adjusted by RV1 and R3 (i.e: the signal is moved up or down) so that the signal which feeds into IC1b produces a stable and reliable pulse at IC1b's output, as shown. This pulse triggers a pulse generater (C6, D6, R6, RV2 and IC1c) whose output pulse width is 60 us ±30 us approximately. (Pulse width adjusted by RV2.)

This gives us two nice pulses indicating the start of the horizontal and vertical sweeps. It is quite likely that the signals available in your monitor or TV will differ from the ones shown here; you will have to try to find signals that you can use either by studying the monitor circuit or prodding around with an oscilloscope. The techniques shown in Figures 6 and 7 are quite generally applicable and should be useful in most cases. In any event, your input signal conditioners should provide the same kind of pulses for the subsequent circuitry to work with. ▶



The Line Uncramper fits neatly into a small metal box attached to the side of the TV set/monitor. The circuit board is supported by the rotary switch only. Connections to the TV set pass through a hole behind the board through the box and TV set cabinet walls.



This short pulse, 'T', serves both as the signal to put a space in the display scan, and also presets IC4 again ready to count the lines in the next character row.

The remaining circuitry is triggered by pulse 'T' to provide the required current to the yoke. First, some words about current sources and sinks, and about how to supply current to the yoke.

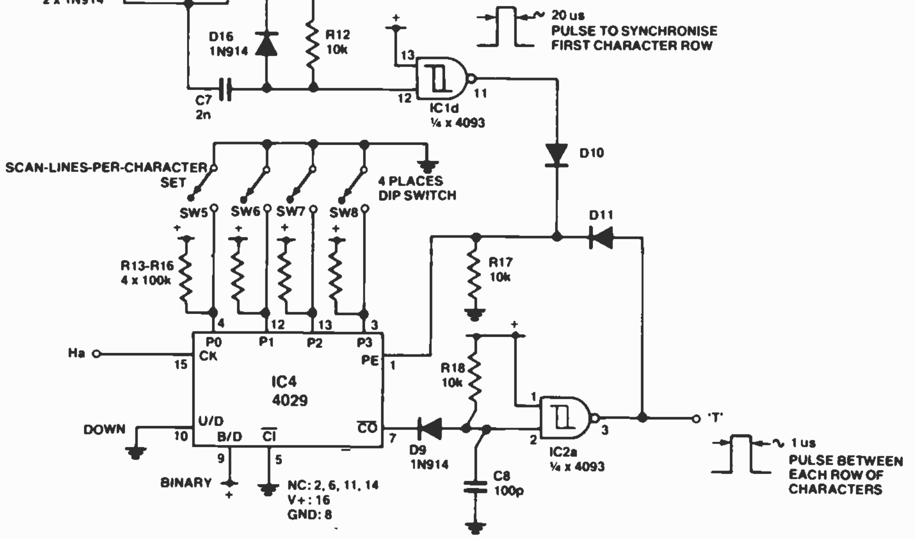


Figure 8: **The timing circuitry is the real heart of the Line Uncramper. It determines where the first space will appear on the screen, and the number of scan lines between subsequent spaces.**

Timing

The purpose of the section shown in Figure 8 is to generate a pulse which indicates that a space is to be made in the screen scanning. It therefore must determine where the first space must occur, and then must repeatedly signal a space after the bottom of each character row. Counter IC3 handles the first part of this task. The pulse from the vertical signal conditioner, 'Va', loads this counter with the number set on DIP switches SW1 to 4. The counter is set up to count in binary, and to count downwards, and it counts the horizontal pulses 'Ha'. When it counts down to zero its 'carry' output synchronises the second counter IC4 by presetting it (via pulse generator C7-D16-R12-IC1d, and then D10) with the value on DIP switches SW5 to 8. This value is the number of lines in a character, which for the Sorcerer is eight or 1000 binary (SW8 open, SW7 to 5 all closed). IC4 also counts downward, and each time it reaches zero its 'carry' output triggers a pulse generator (D9, R18, C8, IC2a).

Current sinks

Some current sources and sinks are shown in Figure 11. In each case the collector lead (marked 'c'), if connected to some other circuit, is able to supply (source) or draw away (sink) a predetermined amount of current, *without* affecting the voltage present on that other circuit (within limits). In Figure 11a, for example, the current drawn away is labelled Ic (collector current). For our approximate purposes, Ic is the same as Ie. Ie in turn is determined by Ohm's Law to be the voltage across resistor R, divided by R. The voltage across R is Vb, less the fixed base-emitter voltage of 0.6 V. Consequently, by setting Vb we can determine Ic. As a formula:

$$Ic = (Vb - 0.6)/R$$

The formula for Figure 11b is the same, while those for Figures 11c and d are similar except that there are two base-emitter voltages, and hence the formula is:

$$Ic = (Vb - 1.2)/R$$

(These circuits, of course, only work for positive values of Ic.

The advantage of the two-transistor design is simply that about 50 to 100 times less base current (Ib) is required to operate it. The limit within which these circuits must operate is that the voltage on the circuit to which the current sink or source is connected must allow at least a certain minimum voltage from collector to ground (11a and 11c) or from collector to the positive supply V+ (11b and 11d). This minimum is Vb−0.4 V approximately.

Uncramper outputs

Now, looking back to Figure 3, what we are attempting to do is to add to, or subtract from, the current flowing in the yoke. Increasing the current through the yoke can be done by feeding more current into point VYx1, or drawing more current away from VYx2. In the first case a current source would be attached to VYx1, while in the second a current sink would be attached to VYx2. The second of these two is not feasible since the small voltage across the current-sensing resistor is insufficient for a current sink to operate.

To reduce the current through the yoke, a current sink may be attached to VYx1, or a current source attached to VYx2. (It may be thought that where the current added or subtracted affects the voltage across the current-sensing resistor, the negative feedback would cause the vertical amplifier to 'correct' for the difference and cancel the desired addition or subtraction. This does not happen because in fact the amount added or subtracted is relatively small, and the indicated feedback loop is not the only feedback path.)

The basic sawtooth current to be added to the yoke is generated by the circuit of Figure 9. Here, the pulse 'T' turns on Q1 and Q2 for a short time. Q2 forces the voltage across C9 to about 2 V. When Q2 turns off, C9 slowly charges via R21 and RV3. The repetition of these events results in a sawtooth waveform at the base of Q3, the height of which is adjusted by RV3. This sawtooth voltage controls a current source (Q3, Q4, R22) which supplies a sawtooth current to the yoke.

That *should* finish the job. Unfortunately, the sawtooth current source is not able to force the drop in current at the end of each tooth to occur fast enough, due to the slowing effect of the yoke's inductance. The result is that the space which should occur *between* scan

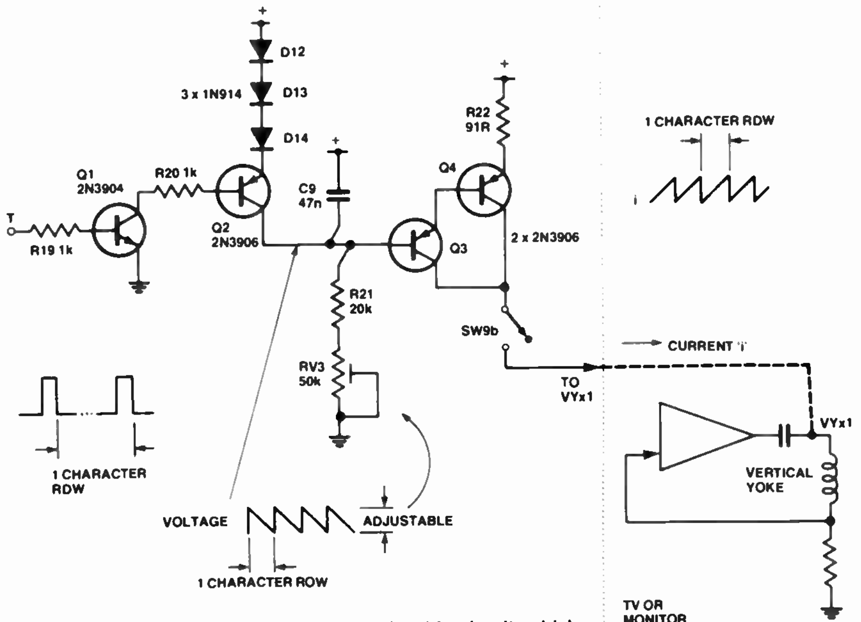
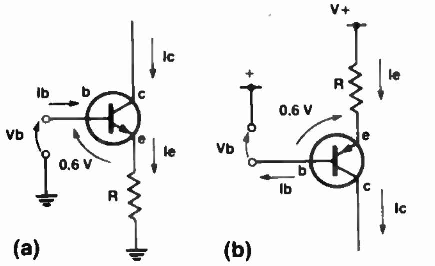
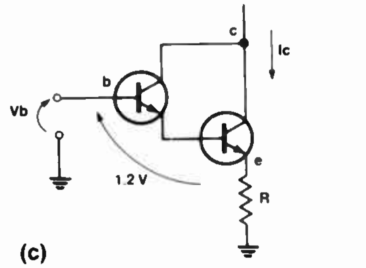


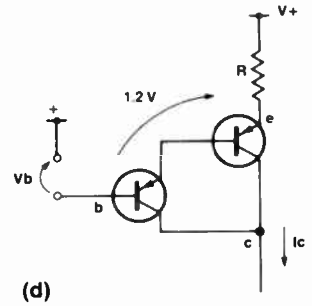
Figure 9: **A sawtooth current is generated by this circuit, which feeds the vertical yoke and causes the vertical squeezing of the scan lines belonging to one character row.**



11a: **NPN current sink.** *11b:* **PNP current source.**



11c: **Two-transistor ('Darlington' arrangement)**



11d: **Darlington current source.**

Figure 11: **Circuitry for current sources and sinks.**

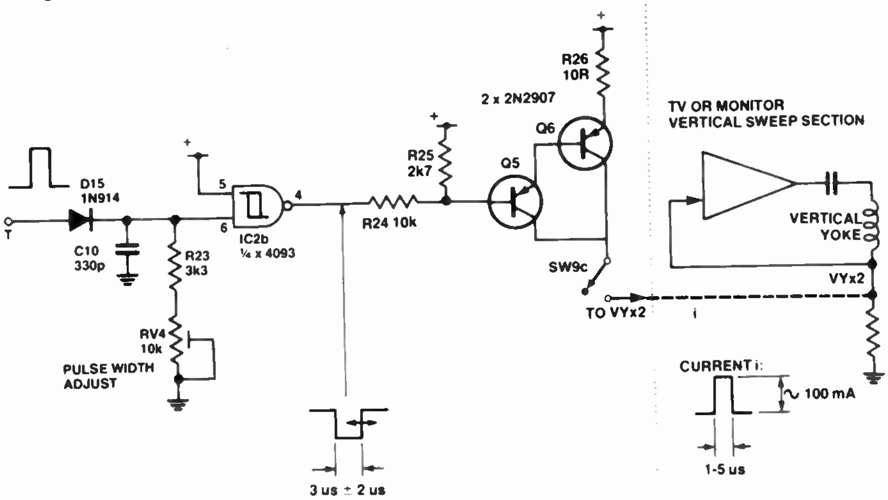


Figure 10: **This circuit sends a pulse to the yoke to make the between-row step in the scan occur sharply.**

lines (in about 5 us or less) actually takes so long (about 50 us) that the top scan line of each character row is on an angle, the left end joined to the upper row of characters, and the right end finally dropped down to join with the lower row of characters. What is needed is to give the yoke a swift kick to cause the step in the vertical sweep to occur smartly. This function is performed by the circuitry of Figure 10.

Pulse 'T' is 'stretched' by D15-C10-R23-RV4 and produces a pulse at the output of IC2b which is 3 us ±2 us long. This pulse is applied to a current source (R24, R25, Q5, Q6), which supplies a pulse of current to VYx2, sharply reducing the voltage across the yoke for a few microseconds.

Switch SW9a-b-c turns on the power to the uncramper, and connects the two current sources to the monitor.

Construction

Any construction method which you have found satisfactory for digital circuitry should be appropriate for the line uncramper. The author's prototype was constructed on Veroboard and mounted in a small metal box attached to the side of the TV set/monitor. Rotary switch SW9 is mounted on the front of the box, and supports the circuit board. Wires into the TV pass through a hole drilled through the box and TV walls.

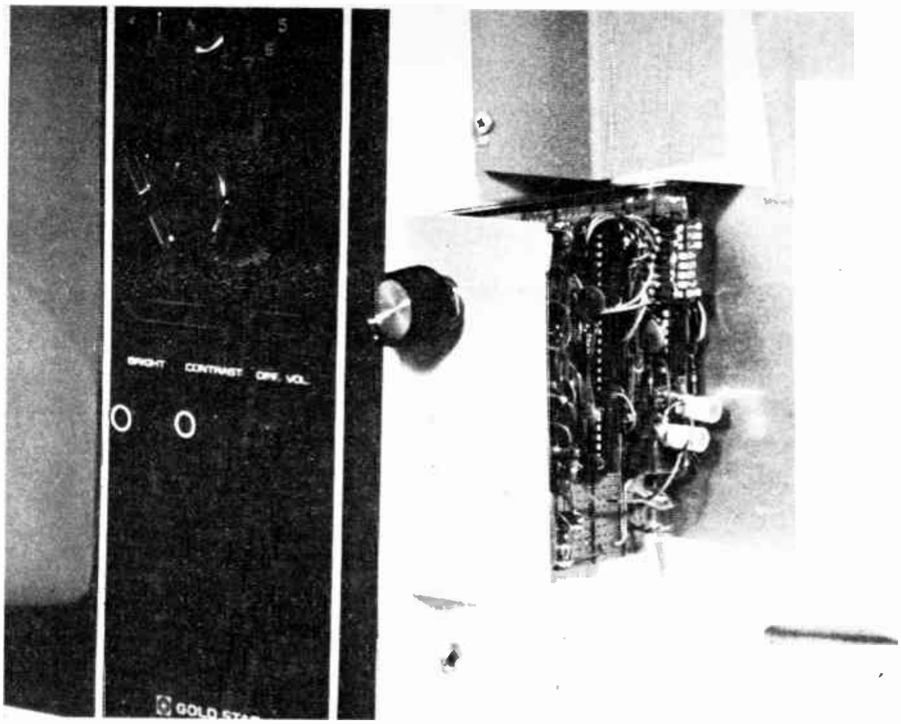
Adjustment

An oscilloscope will probably already have been needed to find the necessary horizontal and vertical signals in the monitor, and will be required now for setting up the uncramper. First set all trimmer potentiometers for the middle of their ranges. Then power up the modified monitor and quickly ensure that the supply voltages on the uncramper are as supposed to be, and that nothing is getting extremely hot, etc. With your computer connected to the monitor, the horizontal and vertical pulse conditioners should be adjusted and verified for proper operation. (It is probably a good idea not to connect the two current sources to the monitor until this stage is reached and the generally satisfactory operation of the uncramper has been checked with the 'scope.)

Next, DIP switches SW5 to 8 can be set for the appropriate number of scan lines per character row. The display ▶

PARTS LIST

Resistors	**all 5%, ⅛ W or greater unless noted**
R1, R4, R19, R20	1k
R2, R5, R6, R11, R12, R17, R18, R24	10k
R3, R7-10, R13-16	100k
R21	20k
R22	91R
R23	3k3
R25	2k7
R26	10R ¼ W
Trimpots	**all linear, miniature, multiturn units are desirable but not necessary.**
RV1, RV4	10k
RV2	100k
RV3	50k
Capacitors	
C1	200u or greater/20V electrolytic
C2	500n non-electrolytic
C3, C5	10u/25 V electrolytic
C4, C8	100p
C6	4n7
C7	2n2
C9	47n
C10	330p
Inductor	
L1	1 mH or greater RF choke
Semiconductors	
D1-D16	1N914 or similar silicon signal diode
Q1	2N3904, BC107, BC547
Q2, Q3, Q4	2N3906, 2N5139, 2N3250
Q5, Q6	2N2907, 2N3645, 2N4143
IC1, IC2	4093
IC3, IC4	4029
Switches	
SW1-8	8-section DIP switch
SW9	3-pole 2-position



The Uncramper switch is positioned in a convenient location for operation alongside the TV set's controls. The upper box attached to the set is the monitor input circuit, allowing the TV set to use the computer video signal directly.

should be looking somewhat reasonable, with horizontal black spaces appearing in the picture. (Increase the brightness so that you can see what the scan lines are doing.) Now switches SW1 to 4 can be changed until the black lines are between rows of characters. Finally, RV3 and RV4 must be alternately adjusted to provide the desired amount of space and proper vertical stepping respectively.

The line uncramper should provide you with a video monitor which is easier to read, and a computer system which is more enjoyable to use. It is a device which may not be necessary with future generations of computers, but in the meantime you can save your eyes . . . and perhaps some enterprising company would like to pick up the idea and provide uncramper kits for the more popular monitors? ●

Super timer —from microseconds to days

Timing long periods, particularly with analogue ciruitry, has always been a problem because of the high leakage characteristics of the timing capacitor. This is no longer true! The XR-2240 IC is a programmable timer capable of producing ultra-long time delays without sacrificing accuracy.

Barry Davis

THIS IC can generate time delays from microseconds up to five days, and with a little ingenuity can generate a delay of a couple of years! A functional block diagram of the IC is shown in Figure 1.

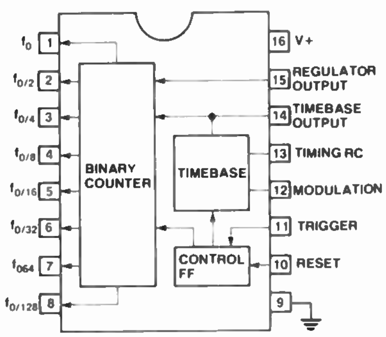


Figure 1. **Functional block diagram of the XR-2240.**

The circuit consists of an internal timebase generator, a programmable 8-bit counter and a control flip-flop. The time delay at the output is set by an external CR network and can be any period from 1.CR to 255.CR. Herein lies the secret. The CR timebase generator can be set to give a very accurate short period, and binary multiples of this short period are then programmed and taken from the output. Each output is capable of sinking approximately 5 mA of load current.

The features of the IC are:

1. Timing from microseconds to days.
2. Programmable delays: 1.CR to 255.CR
3. Wide supply voltage range: 4 V to 15 V
4. TTL compatible inputs and outputs
5. High accuracy: 0.5%
6. Excellent temperature stability
7. Period T = C x R

Circuit operation

The timing cycle is initiated by a positive-going pulse on pin 11. This trigger pulse performs three functions:

1. Activates the timebase generator
2. Enables the counter
3. Sets all counter outputs to the *low* state

The timebase generator produces timing pulses with a period, T, equal to 1.CR. These clock pulses are counted by the binary counter inside the IC and the timing period is complete when a positive-going pulse is applied to pin 10 (i.e: the circuit is reset). In most applications one or more of the output terminals are connected back to the reset input. The circuit will commence timing when the trigger pulse is applied, and automatically reset on the completion of the timing period.

Remember, the outputs are normally high and are set to low when timing is initiated, returning again to the high level on completion of the timing period. ▶

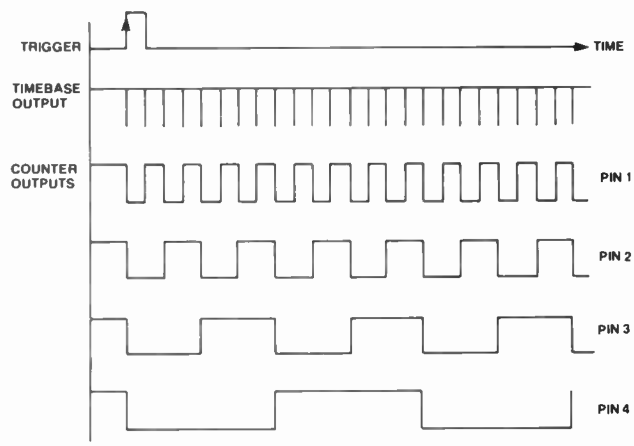


Figure 2. **Output waveforms and timing diagram.**

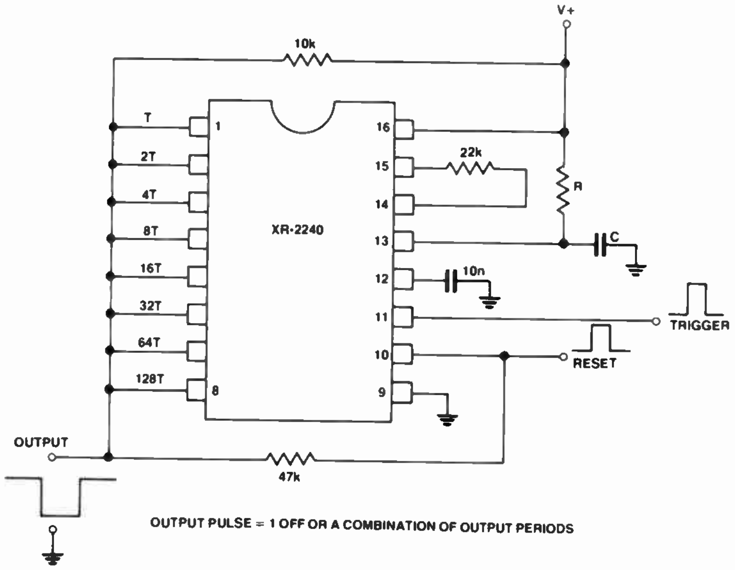


Figure 3. **Connections for a practical circuit.**

Circuit construction

The binary outputs, pins 1 to 8, are open collector and can be connected together to a common pull-up resistor. The output of the timer will be low as long as any one output is low. In this manner the time delays associated with each output can be added by simply connecting them together to a common bus. The outputs can be used individually or wired together.

For example, the output at pin 4 is 8 x CR = 8T. If pins 4 and 3 are connected together the output will become 12 x CR = 12T.

Figure 3 shows the actual connections for a practical circuit. When the power is applied, with no trigger or reset inputs, the circuit sets up to the initial state of all outputs high. Once triggered, the circuit is totally immune to any additional trigger inputs until the timing period is completed, or a reset pulse is applied.

Choice of timing components

Once a signal timing period, T, is established, the output can be determined by 'wiring-in' periods of T following a binary progression. However, the procedure may have to be reversed when a certain accurate output period is required.

For example, if a timing period of 6 hours 30 seconds is required, firstly convert the time to seconds:

$$= 6 \times 60 \times 60 + 30$$
$$= 21630 \text{ seconds.}$$

The maximum number of timing periods available with one IC is $(1+2+4+8+16+32+64+128)T = 255\,T$. Therefore the period of T can be calculated:

$$T = \frac{21630}{255}$$
$$= 84.82 \text{ seconds}$$

With a low-loss capacitor (such as tantalum) as one timing component, R can be calculated. If C = 100 uF:

$$T = CR$$

Therefore:

$$R = \frac{T}{C}$$
$$= \frac{84.82}{100 \times 10^{-6}}$$
$$= 848.24\ k\Omega$$

This can be tailored precisely for very accurate timing with a resistive network or potentiometer, or simply rounded off to 850k.

Figure 4 shows two graphs which will assist you in choosing:

1. The recommended range of timing component values.
2. The time period, (T) up to 100 seconds, to be expected from combinations of C and R values.

An example of output periods to be expected using a 100 uF capacitor (tantalum) and 1M resistor as the timing components is shown in Table 1.

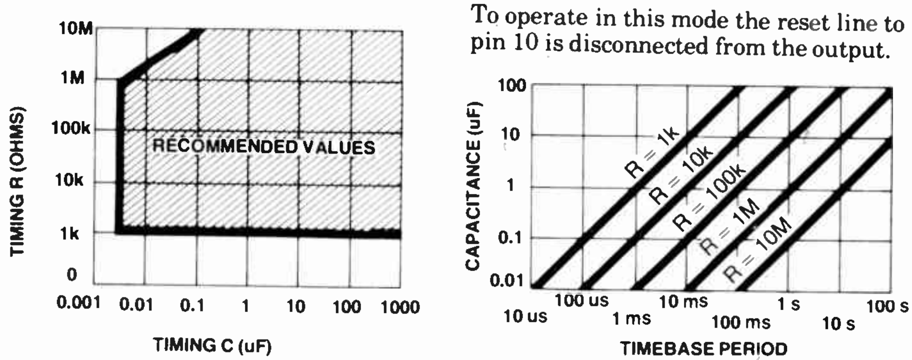
T = CR = 100 µF x 1 M = 100 seconds	
	Period of Output
T	100 secs = 1.7 min
2T	200 secs = 3.3 min
4T	400 secs = 6.7 min
8T	800 secs = 13.3 min
16T	1600 secs = 26.7 min
32T	3200 secs = 53.3 min
64T	6400 secs = 1.8 hours
128T	12800 secs = 3.6 hours
255T	25500 secs = 7.1 hours

Table 1. Example of accurate time available using the XR-2240.

The type of circuit operation discussed to this point has been monostable i.e: the output goes low on triggering, stays low for the timing period and returns to a high level. It will not time again until it is retriggered. An XR-2240 can also be used in a free-running or astable mode.

Astable operation

To operate in this mode the reset line to pin 10 is disconnected from the output.



a) RECOMMENDED VALUES

b) NOMOGRAPH OF CR PERIODS

Figure 4. **Graphs to assist in the choice of values of C and R.**

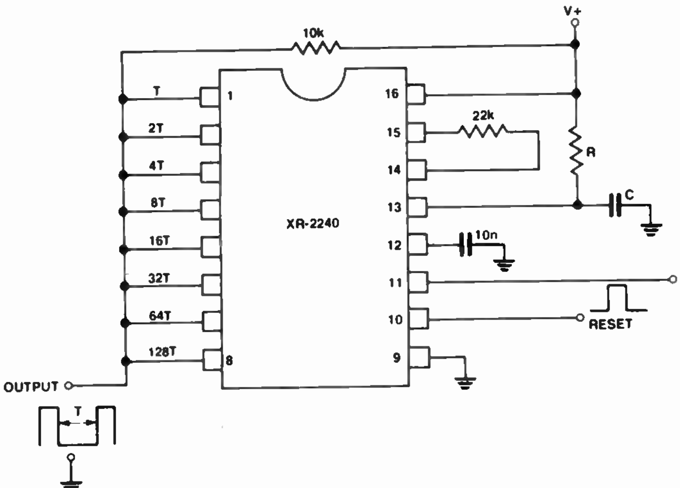


Figure 5. **Astable operation under control of external trigger and reset controls.**

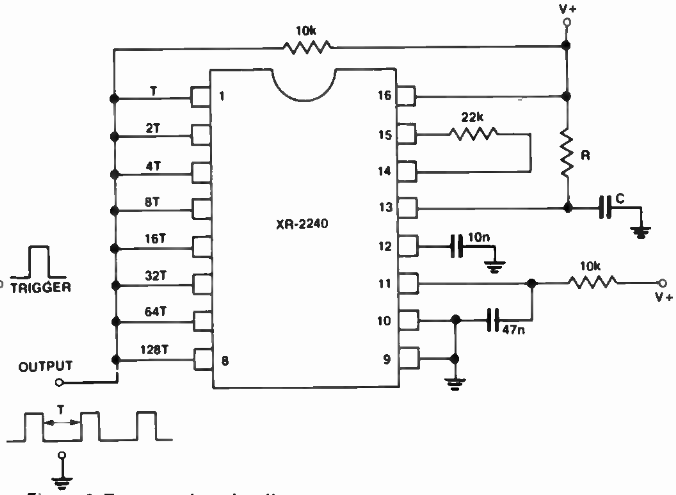


Figure 6. **Free-running circuit.**

Figure 5 shows an astable circuit under the control of the external trigger and reset signals. It will start timing when an external trigger pulse is applied, and will not stop until a reset pulse is applied.

Alternatively, the circuit can be made truly free running. The circuit in Figure 6 self-triggers automatically when the power is switched on, and continues to operate in its free running mode indefinitely.

When the timer is used in this mode, each counter output can be used individually as synchronised oscillators, or they can be connected together to provide complex pulse patterns.

Ultra-long delays

In some applications delays of four days may be required. This is particularly useful in electronic farming for controlling the rate of supplementary feeding. The timing components required can be calculated thus:

4 days = 96 hours
= 5760 minutes
= 345 600 secs.

Maximum number of T combinations = 255

Therefore

$$T = \frac{345600}{255}$$

$= 1355.3$ secs

$= 22.6$ minutes

Incidentally, 20 minutes is about the longest time recommended for 1.CR as anything beyond this suffers from leakage problems.

if $T = CR$

$C = 500$ uF (low leakage)

$$R = \frac{T}{C}$$

$$= \frac{1355.3}{500 \times 10^{-6}}$$

$= 2M7$

Two XR-2240 ICs can be cascaded to generate extremely long time delays. When used in this format the reset and trigger terminals of the ICs are tied together and the timebase of unit 2 disabled as shown in Figure 7.

The output is normally high. When a positive-going trigger pulse is applied the output goes low and stays in the low state for $(256)^2 = 65\,536$ periods of the timebase oscillator. Therefore the total timing period of two cascaded units can be from 256.CR to 65 536.CR. The output is available in 256 discrete steps by selectively connecting one or a combination of the outputs from unit 2 to the output bus.

With T = 20 minutes an example of an ultra-long delay can be given.

CR = T = 20 mins.

65 536T = 1 310 720 min
= 21 845 hours
= 910 days
= 2.5 years!

This article highlights the use of an XR-2240 as a precision timer. Other application suggestions are:

1. Sequential timing
2. Binary pattern generation
3. Frequency synthesis
4. Pulse counting or summing
5. A/D conversion
6. Digital sample and hold

Further information on the IC can be obtained from Exar Integrated Systems or their agents (Total Electronics).

This article was made possible by the courtesy of Exar Integrated Systems. Data was taken from their publication 'XR-2240/2340 Programmable Timer Counter'. ●

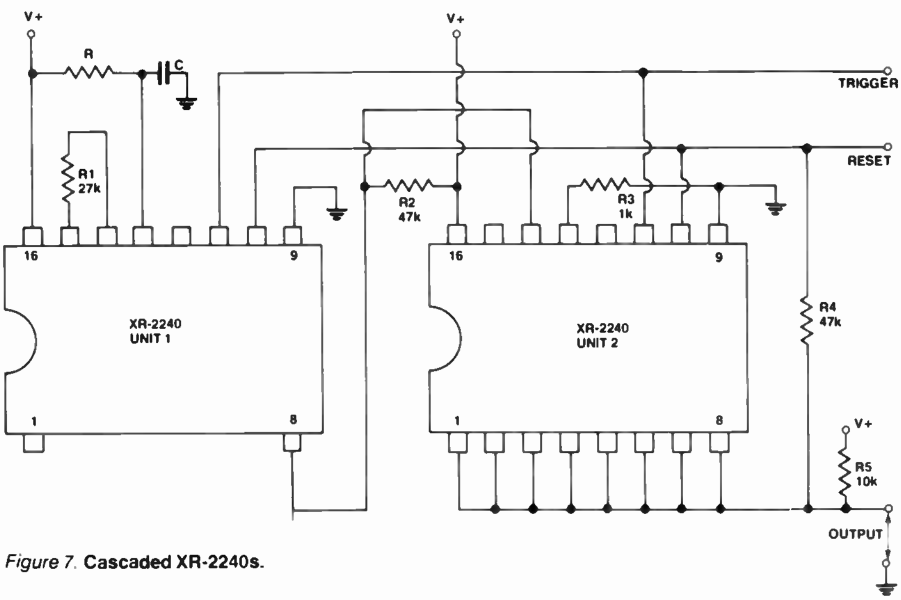


Figure 7. **Cascaded XR-2240s.**

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Wideband antenna baluns

Few antennas are co-operative when it comes to feedpoint impedance. Matching them up or down to the required 50 ohms (occasionally 75 ohms) often proves awkward. Roger Harrison shows the easy way round.

IN ANY COMMUNICATIONS system the weakest link in the chain is always the antenna system. Even if the best antenna to suit the particular circumstances is chosen, it is necessary to provide efficient transfer of power to or from the antenna feedpoint. However, the antenna feedpoint impedance is not always conveniently the same as that of the feedline, the transmitter output impedance or the receiver input impedance. The latter are usually either 50 or 75 ohms to match the coaxial transmission lines normally employed. Coaxial transmission lines are, by nature unbalanced electrically, whereas many types of antennas have balanced feedpoints and thus require a balanced transmission line. To correct this a 'balun' is necessary – the word being derived from 'balanced – to – unbalanced'. Originally, the term applied to a device which only involved changing from the balanced to unbalanced condition without a change in impedance. Where a change in impedance is necessary the term transformer is, strictly speaking, the correct term.

There are a wide variety of methods of making antenna baluns and impedance matching devices. However, most techniques are suitable for use on a single, narrow frequency band. The most versatile technique, which results in baluns that may be used over a wide frequency range, employs sections of made-up transmission line wound on a ferrite core, usually in the form of a toroid or some other convenient shape. Sections of transmission line are wound together and connected in a series or parellel combination to effect the desired balanced-to-unbalanced transformation and/or an impedance transformation. Winding the transmission line sections on a ferrite core increases the inductance of the length of transmission line used. This article shows how to make the most useful types using locally available components.

High Power Transmitting Baluns

The following baluns to be described all employ toroids and are for use in transmitting applications at power levels up to 200 W, and up to 1 kW, CW or PEP output. Only two toroids are specified, both from the Neosid range. For applications up to 200 W, the toroid type 4328R/2/F14A/EC is employed, while that for powers up to 1 kW employ the type 4324R/3/F14A. The first has an outside diameter of 25.4mm, an internal diameter of 19.05 mm and a thickness of 9.52 mm. It is coated in an enamel paint. The second, and larger toroid has an outside diameter of 38.1 mm, an inside diameter of 25.4 mm, is 19.05 mm thick and is uncoated. Both have bevelled edges and will not nick or cut the insulation of wire wound around them.

The smaller toroid may be used for purely receiving applications if desired, particularly for the higher transformation ratios of 9:1 and 16:1.

The baluns described are suitable for use from 2 MHz to 30 MHz and in some cases a wider range.

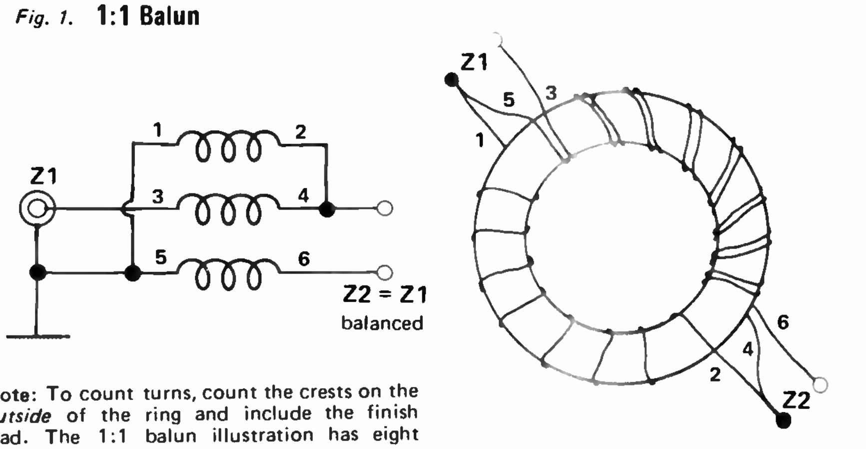
1:1 Balun

The circuit and connections for a 1:1 balun are given in Figure 1. This uses a bifilar winding wound around half the core and a single winding around the other half as illustrated. One wire in the bifilar winding is connected in series with the single winding. This balun is some times described with a trifilar winding but balance and bandwidth are not as good as with the method described.

This type of construction exhibits excellent balance at the Z2 termination and operates over a bandwidth of 1 MHz to 30 MHz in 50 ohm systems for assemblies on either toroid. The bandwidth is considerably better for 75 ohm systems constructed on either toroid, extending to 60 MHz for the 1 kW assembly but only 40 MHz for the 200 W assembly.

The exact gauge of the wire used is not very critical, a latitude of plus or minus one gauge being tolerated. When winding on the heavy gauge wires, a ▶

Fig. 1. **1:1 Balun**



Note: To count turns, count the crests on the *outside* of the ring and include the finish lead. The 1:1 balun illustration has eight turns on each winding.

	200W	1KW
Winding details:	8 turns, 20 or 22 SWG enamelled wire for both bifilar and single windings.	6 turns, 18 SWG enamelled wire for both windings.
Response		
50 – 50 ohms:	1 – 30 MHz	1 – 30 MHz
75 – 75 ohms:	1 – 40 MHz	1 – 40 MHz

neat winding can be obtained by carefully forming each bend, holding the already wound part firmly against the core.

Some confusion arises when counting turns of a winding on a toroid. To count turns, count the 'crests' on the outside of the ring and include the finish lead.

The assembly illustrated in Figure 1 has eight turns on each winding.

4:1 Baluns

An unisolated balun has a dc connection between input and output, the isolated type is a true transformer.

Figures 2 and 3 illustrate unisolated types. A 4:1 isolating balun is in Fig. 4.

A single bifilar winding is used. The wires may be twisted together lightly or wound together around the core as illustrated. The latter method is preferable. Identify the ends and connect in series as shown. The bifilar winding is spread around most of the circumference. For the larger toroid, it is difficult to spread six turns around the circumference so the winding is distributed around about two-thirds of the circumference of the toroid.

Balance of the high impedance is excellent. Bandwidth for both toroids is best for Z1 of 50 ohms. The 1 kW assembly in this case has the best bandwidth for either 50 or 75 ohm systems,

The same assembly can be connected as an unbalanced transformer. The arrangement is shown in Fig.3. Bandwidth is limited in this application but most vertical or loaded vertical antenna systems are used on the lower bands, below 15 MHz in any case. Best bandwidth is obtained for transformation from about 19 ohms to 75 ohm systems.

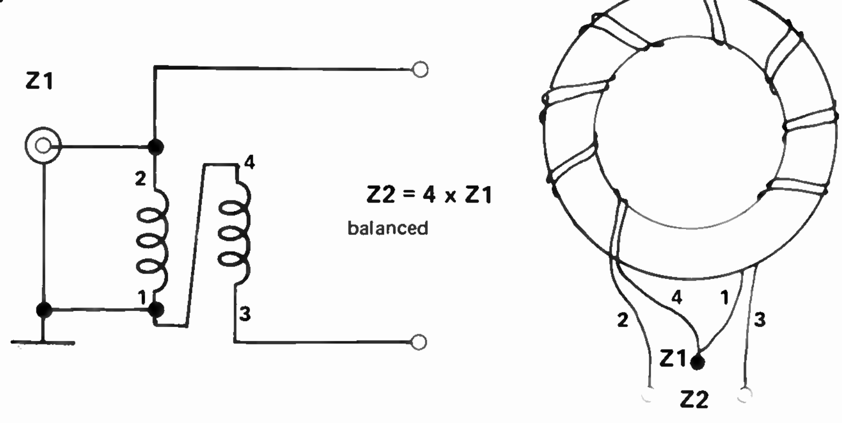
The isolated type is illustrated in Figure 4. This consists of a trifilar winding having two of the wires connected in series for the high impedance winding, the third wire being the low impedance, unbalanced winding.

The 1 kW assembly has the best bandwidth, when matching to 50 ohm systems. However, the full HF range is covered by both assemblies.

9:1 Balun

Baluns providing high impedance ratios are rarely described. Recently, wideband transformers for use in HF transistor linear power amplifiers, having impedance ratios as high as 36:1, have been described, but the techniques used are not generally adaptable to the applications considered here.

Fig. 2. **4:1 Balun**

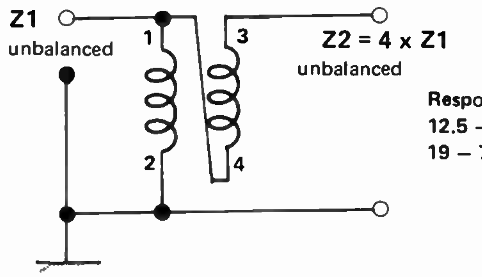


	200W	1KW
Winding details:	10 turns bifilar, 20 or 22 SWG enamelled wire, wound around most of circumference.	6 turns bifilar, 16 or 18 SWG enamelled wire, wound around most of circumference.
Response		
50 – 200 ohms:	1 – 75 MHz	1 – 60 MHz
75 – 300 ohms:	1 – 60 MHz	1.8 – 30 MHz

Many types of HF beam antennae have high impedance feedpoints, such as the Lazy-H, Sterba curtain, V-beams and Rhombics, therefore presenting a matching problem that is usually solved by using resonant matching devices or 'match-boxes' involving tuned transformers. The wide bandwidth advantage of the V-beam and the Rhombic is compromised by such devices and a wideband balun provides a much better solution to the problem of matching the balanced, high impedance feedpoint to the unbalanced, low impedance transceiver antenna terminal. You don't have to tune up each time you change bands. Conversely, you lose the harmonic and spurious suppression advantages of the tuned matching system. However, it is always good practice to insert a low-pass filter in the transmission line immediately following the transceiver, regardless of the matching system used.

For antennas having a feedpoint impedance close to 600 ohms the balun illustrated in Figure 5 is applicable. It consists of three separate bifilar windings wound on the ferrite toroid as shown. One wire from each winding is connected so that each is in series, this forming the balanced high impedance winding. The remaining wires from each winding are all connected in

4:1 Unbalanced transformer



	200 W	1 kW
Response		
12.5 – 50 ohms:	1 – 15 MHz	1 – 18 MHz
19 – 75 ohms:	1.8 – 28 MHz	1 – 30 MHz

Construction as per Figure 2., connections as shown at left.

Fig.3. Circuit for 4:1 unbalanced to unbalanced transformer. Construction is similar to that in Fig.2. This type of transformer is suitable for matching to low impedance unbalanced antennas.

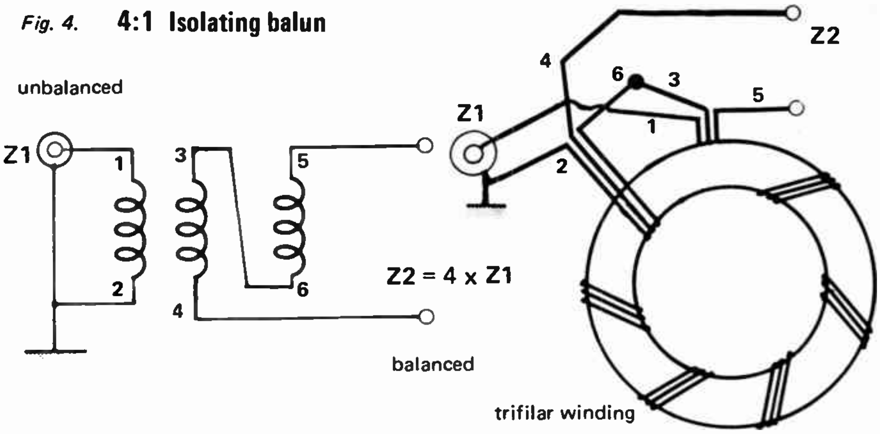
parallel for the unbalanced low impedance winding.

A single pair stripped from a length of suitable ribbon cable, or rainbow cable as it is also called, may be used for the windings on the small toroid specified for the 200 W assembly. For the1 kW assembly on the larger toroid, a type of 'figure – 8' flex that is sold as speaker cable is conveniently used. It has one lead marked with a dark-coloured stripe which helps to identify the separate wires in each bifilar winding. The different coloured insulation of the ribbon cable wires serves the same purpose. The figure-8 flex sold as 240 V lamp and appliance cord is too large to be used here.

The assembly illustrated will cover the range 3 MHz to 25 MHz with eight turns per winding on the small toroid and six turns per winding on the larger toroid. More turns per winding are required if the balun is to be used at frequencies lower than 3 MHz. However, as the assembly only has a bandwidth of about 8:1 the upper frequency is then limited to about 15 MHz. The number of turns required barely fits on each core in this case.

Although the balun described is specified for matching 600 ohms to 75 ohms it can also be used to match antennas having impedances close to 450 or 500 ohms, to 50 ohms. The upper frequency limit is then reduced to about 20 MHz in this case. A slight impedance mismatch is readily tolerated by most equipment and there is little to be gained in trying to get an exact match.

Fig. 4. **4:1 Isolating balun**



	200W	1KW
Winding details.	8 turns trifilar, 20 or 22 SWG enamelled wound around most of circumference.	6 turns trifilar, 16 or 18 SWG enamelled wound around most of cumference.
Response		
50 – 200 ohms:	1.8 to 28 MHz	1.8 to 75 MHz
75 – 300 ohms:	3.0 to 35 MHz	2.5 to 30 MHz

16:1

For antennas having feedpoint impedances in the vicinity of 800 ohms the balun illustrated in Figure 6 is applicable for matching to 50 ohms. It is constructed in a similar manner to the previous one. Four bifilar windings are wound on the core. Four wires, one from each winding are connected in series to form the high impedance balanced winding. The other four wires are connected in parallel for the low impedance unbalanced winding.

A single pair stripped from ribbon cable is also conveniently used for this balun on the small toroid, while figure-8 flex is convenient for the larger toroid, as discussed for the 9:1 balun.

This balun covers 3 to 25 MHz for assemblies wound on either toroid. Sufficient turns will not fit on the smaller toroid if you wish to go below 3 MHz. For purely receiving applications, use light gauge enamelled copper wire – such as 28 or 30 swg, and wind eight or nine turns per winding of bifilar pairs on the small toroid. As receivers are usually fairly tolerant of some degree of mismatch, a balun constructed in this manner may be used over the range 1 MHz to about 25 MHz.

The larger toroid will fit sufficient turns to cover the range down to 1.8 MHz but, as for the 9:1 balun, the upper frequency limit is about 15 MHz.

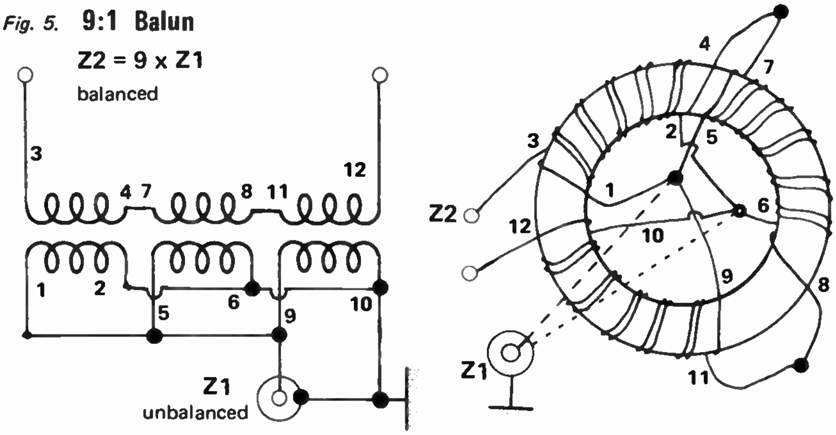
TOROIDS

200W: 4328R/2/F14A/EC
Size: 25.4 mm x 19.05 mm x 9.52 mm (od, id, depth). Suitable up to 200 W and for receiving applications.

1 kW: 4324R/3/F14A
Size: 38.1 mm x 25.44 mm x 19.05 mm (od, id, depth)

The toroids specified are manufactured by Neosid (Aust.) Pty Ltd, 23-25 Percival St, Lilyfield NSW 2040; phone (02) 660-4566.

Fig. 5. **9:1 Balun**

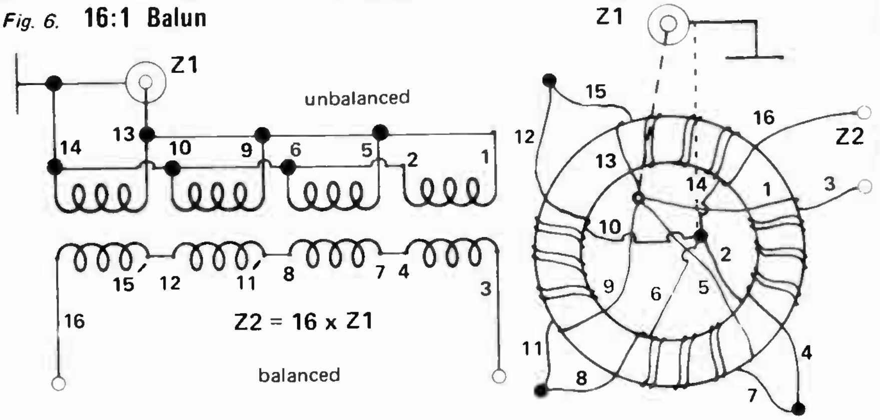


	200W	1KW
Winding details:	* For use between 3 – 25 MHz	
	Three bilfilar windings 8 turns each, 12/0076 PVC covered hook up wire.	Three bilfilar windings 6 turns each, 10/010 PVC covered hook up wire or "figure 8" flex.
	* For use between 1.5 – 15 MHz	
	10 turns per winding as above.	8 turns per winding as above.

Encapsulating Balun Assemblies

Where necessary, to protect them from the weather, the balun assemblies may need to be encapsulated. A suitable mould may be fashioned from stiff paper or cardboard and the balun assembly potted in a suitable epoxy compound. The input and output terminations may be potted along with them, ensuring that no short circuits are possible in the process. Alternatively, they may be sealed in a suitable plastic moulding such as those produced for 240 V electric cable conduit. These are available as T-junctions etc, and are generally obtainable from hardware stores in sizes suitable to contain the toroids specified. A complete balun and dipole feedpoint termination can be encapsulated in an appropriate fitting. A suitable coax socket and eye-bolts to take the antenna strain are readily included also. The exact construction is left up to the individual to suit the circumstances. A little imagination goes a long way. ●

Fig. 6. **16:1 Balun**



Winding details:	200W	1KW
	* For use between 3 – 25 MHz	
	Four bifilar windings 6 turns each, 12/0076 PVC covered hook up wire.	Four bifilar windings 4 turns each, 10/010 PVC covered hook-up wire or "figure 8" flex.
	* For use between 1.8 – 15 MHz	
	Sufficient turns will not fit on the toroid. See text.	6 turns per winding as above.

Data sheets explained

The data sheets which we publish regularly are very popular, but from time to time we receive requests for a fairly simple explanation of the terms and abbreviations which one finds in semiconductor device data sheets. This article has been prepared to satisfy your requests. **by Brian Dance**

THE INFORMATION contained in semiconductor device data sheets is often grossly misunderstood. Great care must be taken to ensure that the exact meaning of a term or abbreviation is clear. As an example, we can quote the following conversation which actually occurred between two people who should both have known better.

A representative of a semiconductor distributor was showing data on a new power device to a lecturer. The lecturer said that the device data was wrong, since the maximum collector current was quoted as 12A and the maximum collector-emitter voltage (V_{CEO}) as 80V; this is a power level of 12 x 80 = 960W, but the maximum permissible dissipation quoted in the data sheet is only 90W. The representative could provide no answer!

The data was, of course, perfectly correct. The problem arose because neither of the people concerned had appreciated the exact meaning of V_{CEO} which signifies the collector-emitter voltage *with the base open circuited*. Under these conditions (with zero base current) the collector current will be very small and the power dissipation in the transistor will also be quite small. Thus there is a great deal of difference between V_{CE} (the collector-emitter voltage under any conditions) and V_{CEO} (the collector-emitter voltage with the base open circuited). If still more information is required, one must look into the SOAR (Safe Operating ARea) graph to ascertain the regions of the collector voltage/collector current curve where the device can be safely operated for limited or unlimited times.

This is a very simple example of the pitfalls one can encounter if ones does not really understand the *exact* meanings of the terms and abbreviations used in data sheets. Such misunderstandings are very common, but not (we hope!) amongst the devices covered in our data sheets, since it is equally important that our readers understand the exact meanings of abbreviations used in data sheets on relatively simple devices such as ordinary diodes and transistors.

LETTER SYMBOLS

Three of the most important symbols used in semiconductor device data sheets are V, I and P for voltage, current and power respectively. Various subscripts are added to these three letters to indicate the electrode(s) to which the symbol is being applied and possibly certain circuit conditions. Some of the most commonly used subscripts are listed below.

A	anode
AV	average
B	base
BO	breakover
BR	breakdown
C	collector
D	drain or delay
E	emitter
F	forward
G	gate
H	holding
I	input
J	junction
K	cathode
M	peak value of a quantity
O	open circuit or output
R	reverse or repetitive
S	source, short circuit, series or shield
T	in the on state (that is, triggered)
W	working
X	specified circuit
Z	regulator impedance

CAPITALS AND LOWER CASE Both the quantity being shown and the subscript may appear as a capital or lower case letter in order to differentiate between instantaneous and rms values. The basic rules are given in the following table:

	MEANING
Capital quantity symbol plus capital subscript [V,I,P] + [C,E,B]	The steady current (no signal) value. The subscript (AV) may be added to indicate the total average value with signal or (M) for the total peak value.
Capital quantity symbol plus lower case subscript [V,I,P] + [c,e,b]	The rms value of the alternating signal component. The subscript (av) may be added to indicate the average value of the varying signal component or (m) to indicate the peak value of this component.
Lower case quantity symbol plus capital subscript [v,i,p] + [C,E,B]	The instantaneous total value of the quantity concerned.
Lower case quantity symbol plus lower case subscript [v,i,p] + [c,e,b]	The instantaneous value of the varying signal component.

Thus i_E is the instantaneous value of the total emitter current, i_e the instantaneous value of the alternating component of the emitter current, and $I_{E(AV)}$ the average (dc) value of the total emitter current. Other subscripts can be used in a similar way, I_F being the forward dc current with no signal, i_F the instantaneous forward current and I_{FM} the peak forward current.

ORDER OF SUBSCRIPTS

In most cases more than one subscript is needed; the subscripts are usually placed in a definite order governed by the following rules:
The first subscript indicates the electrode at which the current or voltage is measured.
The second subscript denotes the reference terminal or circuit mode. (This subscript is often omitted if it is felt no ambiguity will arise.)
The letter O may be used as a third subscript to show that the electrode not indicated by any previous subscript is open circuited. Similarly the letter S can be used as a third subscript to show the third electrode is shorted to the reference electrode of the second subscript, whilst the letter R as a third subscript indicates that a specified resistance is connected between the third electrode and the reference electrode.
The supply voltage to a collector is indicated as V_{CC}, the second suffix being a repetition of the first in the case of supply voltages. Similarly, one often meets the symbol V_{DD} for the positive supply to a CMOS (or COS/MOS) device, this being the supply to the drain. The negative supply to CMOS devices is normally represented by the symbol V_{SS}.

It should now be clear why V_{CEO} is the steady collector-emitter voltage with the base open circuited. Similarly I_{CER} is the collector cut off current with a specified resistance between the base and emitter. It is current with the base and emitter joined, since either the base or emitter can be used as the reference electrode without any change when they are joined.

The parameters of individual devices vary from one device to another of the same type number. The typical value of a parameter such as transistor current gain is often quoted in data sheets by the abbreviation 'typ' after the quantity, but minimum and maximum values are also often quoted. In economical devices no maximum and minimum values may be quoted. In the case of breakdown voltages, the minimum value applicable to any device of that type number is usually quoted so that the circuit designer knows that he can apply that value of voltage without danger of the device junction breaking down.

The above discussion gives the general principles of the way in which the symbols for various parameters are chosen. It is not complete, since we have not yet covered such items as current gain of a transistor or thermal characteristics of a device. However, these and other quantities will be covered in the following tables.

THERMAL CHARACTERISTICS

The symbols used for the following thermal quantities apply to all types of semiconductor device.

P_{tot}	total power dissipated within the device
T_{amb}	ambient temperature
T_c	temperature of the case of the device
T_j	temperature of the junction in the semiconductor material
T_{mb}	temperature of the mounting base of the device ($= T_c$)
T_{stg}	storage temperature
θ_h	thermal resistance of heat sink. (Units. ^{o}C/W)
θ_i	contact thermal resistance between the case of the device and the heat sink
$\theta_{j\text{-}amb}$	junction to ambient thermal resistance
$\theta_{j\text{-}c}$	junction to case thermal resistance

SYMBOLS USED MAINLY WITH DIODES

C_d	diode capacitance with reverse bias
C_f	diode capacitance with forward bias
C_j	capacitance of the junction itself
C_{min}	minimum capacitance (which occurs at the rated breakdown voltage)
C_o	diode capacitance at zero bias
f_{co}	cut off frequency of a varactor diode
I_F	total dc forward current
i_F	instantaneous forward current
$I_{F(AV)}$	average forward current
I_{FM}	peak forward current
I_{FRM}	repetitive peak forward current
I_{FSM}	non-repetitive peak forward current occurring under surge conditions
I_R	continuous reverse leakage current
i_R	instantaneous reverse leakage current
I_{RRM}	repetitive peak reverse current
I_{RSM}	non-repetitive peak reverse current
I_Z	zener diode continuous operating current
I_{ZM}	zener diode peak current
t_{on}	turn on time
t_{off}	turn off time
t_r	rise time
t_{rr}	reverse recovery time
t_s	storage time
V_F	steady forward voltage
v_F	instantaneous forward voltage
V_R	steady reverse voltage
v_R	instantaneous value of the reverse voltage
V_{RM}	peak reverse voltage
V_{RRM}	repetitive peak reverse voltage
V_{RSM}	non-repetitive peak reverse voltage (on surges)
V_Z	zener diode working voltage

SYMBOLS USED MAINLY WITH TRANSISTORS

C_{ob}	transistor output capacitance in the grounded base circuit
C_{oe}	transistor output capacitance in the grounded emitter circuit
f_T	transition frequency or gain-bandwidth product in common emitter circuit
h_{FE} (h_{FB}, h_{FC})	current gain in the grounded emitter circuit (or in the grounded base or grounded collector circuit).
h_{fe}	the increase in collector current divided by the small increase in the base current which produces it. (Small signal current gain.)
I_B, I_C or I_E	the steady base, collector or emitter current.
$I_{B(AV)}$, $I_{C(AV)}$ or $I_{E(AV)}$	the average value of the base, collector or emitter current.
I_{CEX}	collector cut off current in a specified circuit
I_{CM}, I_{BM} or I_{EM}	peak value of collector, base or emitter current
I_b, I_c or I_e	rms value of the alternating component of the current

I_{bm}, I_{cm} or I_{em} — peak value of the alternating component of the current
i_C, i_B or i_E — instantaneous value of the total current
i_c, i_b or i_e — instantaneous value of the alternating component of the current
I_{CBO} — collector cut off current with the emitter open circuited
I_{CBS} or I_{CES} — collector cut off current with emitter shorted to the base
I_{CEO} — collector cut off current with the base open circuited
I_{CER} — collector cut off current with a specified value of resistance between the base and the emitter
I_{EBO} — emitter cut off current with the collector open circuited
$V_{BE(SAT)}$ — base-emitter saturation voltage
$V_{(BR)}$ — breakdown voltage
$V_{(BR)CBO}$ — collector to base breakdown voltage with emitter open circuited
$V_{(BR)CEO}$ — collector to emitter breakdown voltage with base open circuited
V_{CB} — collector-base voltage
V_{CBO} — collector to base voltage with emitter open circuited
V_{CC} — collector supply voltage
V_{CE} — collector to emitter voltage
V_{CEO} — collector to emitter voltage with base open circuited
V_{ce} — collector to emitter rms voltage
$V_{CE(SAT)}$ — collector to emitter saturation voltage
V_{EB} — emitter-base voltage
V_{EBO} — emitter-base voltage with collector open circuited
V_{eb} — emitter-base rms voltage

SYMBOLS USED MAINLY WITH FETS

I_D — steady value of the drain current
I_{DSS} — steady value of the drain current with the gate connected to the source
I_{DM} — peak drain current
I_G — steady gate current
I_S — steady source current
r_{DS} — drain to source (or channel) resistance
V_{DS} — steady drain to source voltage
V_{GS} — steady gate to source voltage

SYMBOLS USED MAINLY WITH THYRISTORS

I_{FRM} — repetitive peak forward current
I_{FSM} — non-repetitive peak (surge) current
I_{GD} — gate current which does not trigger the device
I_{GT} — gate trigger current
I_{GQ} — gate turn off current
I_H — holding current required to maintain conduction
I_R — steady reverse leakage current
I_{RG} — reverse gate current
I_{RRM} — repetitive peak reverse current
I_{RSM} — non-repetitive peak reverse current (in surge conditions)
I_T — steady anode-cathode 'ON' state current
P_G — gate power
t_{gt} — gate controlled turn-on time
t_{gq} — gate controlled turn-off time
$V_{(BO)}$ — breakover voltage
V_D — continuous off state voltage
V_{FG} — forward gate voltage
V_{GT} — gate trigger voltage
V_R — steady reverse voltage

OPERATIONAL AMPLIFIER TERMS

Bandwidth, Δf. The frequency at which the gain falls by a factor of 0.7 relative to the gain at low frequencies.
Common mode rejection ratio, CMMR. The gain when a signal is applied to one of the inputs of the amplifier divided by the gain when the signal is applied to both the inverting and non-inverting inputs. It is usually expressed in dB.
Frequency compensation. An operational amplifier requires a capacitor to enable it to be used in circuits which are stable over a wide frequency range. Internally compensated operational amplifiers have this capacitor fabricated on the silicon chip, but an external capacitor must be used with other types of operational amplifier which do not contain an internal capacitor
Input bias current, I_{bias}. The mean value of the currents at the two inputs of an operational amplifier.
Input offset current, I_{OS}. The difference in the two currents to the inputs of an operational amplifier. Normally much smaller than the input bias current.
Input offset voltage, V_{OS}. The voltage which must be applied between the two input terminals through equal resistors to obtain zero voltage at the output.
Open loop voltage gain, A_{VOL}. The amplifier gain with no feedback applied.
Output resistance, R_O. The small signal resistance seen at the output when the output voltage is near zero.

VOLTAGE REGULATOR TERMS

Dropout voltage, V_{DO}. When the difference between the input and output voltages falls down below the dropout voltage, the device ceases to provide regulation.
Foldback current limiting. In regulators with foldback current limiting, the current will 'fold back' to a fairly small value when the output is shorted.
Line regulation. The change in the output voltage for a specified change in the input voltage.
Load regulation. The change in output voltage for a change in the load current at a constant chip temperature.
Quiescent current, I_Q. The current taken by the regulator device when it is not delivering any output current.
Ripple rejection. The ratio of the peak-to-peak ripple at the input of the regulator to that at the output. Normally expressed in dB.

MONOLITHIC TIMER TERMS

Comparator input current. The mean current flowing in the comparator input connection during a timing cycle.
Timing capacitor, C_t. This capacitor is normally connected between the comparator input and ground. The time taken for it to charge controls the delay time.
Timing resistor, R_t. This is the resistor through which the timing capacitor charges.
Trigger current. The current flowing in the trigger input connection, at the specified trigger voltage.
Trigger voltage. The voltage required at the trigger pin to initiate a timing cycle.

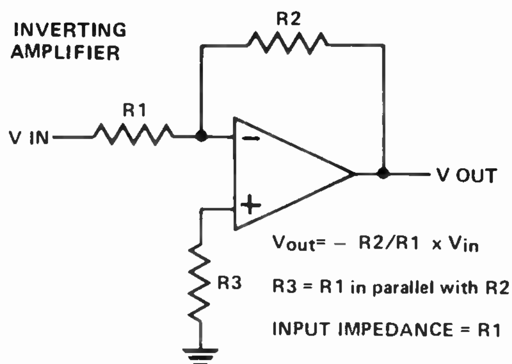
Conclusions

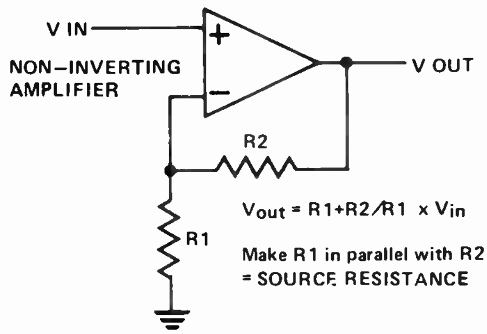
Data sheets must be used intelligently and with much thought. Information on the conditions under which an entry in the data sheet is applicable is often stated in small print, but is of great importance. Data should always be thoroughly studied before a device is used for the first time, only then will you be able to fully understand the potential applications of the device.

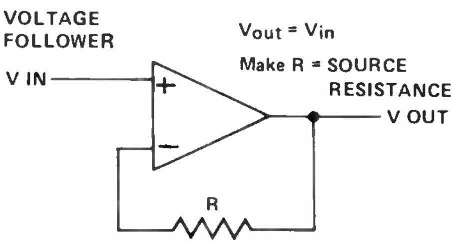
Operational amplifier survey

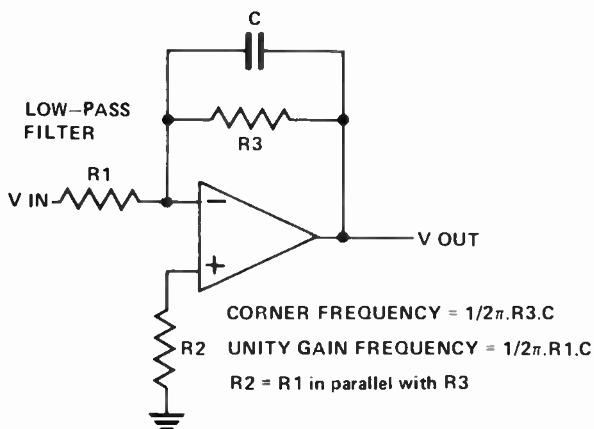
Op amp types	Input offset voltage mV	Input bias current nA	Type of input structure	Band-width MHz	Slew rate V/ns	Voltage gain dB	Maximum supply voltage V	CMRR dB	Qty	COMMENTS
709	2	300	NPN	1	0.25	90	± 18	90	S	Needs frequency compensation
307	2	70	NPN	1	0.25	100	± 18	90	S	Internal frequency compensation
301	2	70	NPN	10	0.5	100	± 18	90	S	Needs frequency compensation
741	2	80	NPN	1	0.5	106	± 18	90	S	Internal frequency compensation
748	1	120	NPN	10	0.5	103	± 22	90	S	A decompensated 741
308	2	1.5	NPN	3	0.5	110	± 18	100	S	Low supply current drain 0.3 mA Needs frequency compensation Very low differential input voltage range
318	4	150	NPN	15	50	106	± 20	100	S	Very low differential input voltage range. Sometimes needs frequency compensation
747	2	80	NPN	1	0.5	106	± 18	90	D	Internal frequency compensation
1458	1	80	NPN	1	0.8	103	± 18	90	D	Internal frequency compensation
4136	0.5	40	PNP	3	1.0	110	± 18	100	D	Low noise
3900 3401	Current inputs	30	Current sinks	2.5	0.5 20	70	± 18	–	Q	Current balancing amplifier
324	2	45	PNP	1	0.5	100	± 30	70	Q	Ground sensing inputs Output voltage can go to ground Low power. 0.8 mA drain oer IC
3403	2	150	PNP	1	1.2	100	± 36	90	Q	Ground sensing inputs Class AB output Output voltage can go to ground Low power 3 mA drain per IC
348	1	30	NPN	1	0.5	103	± 18	90	Q	Low power 2.4 mA drain oer IC Class AB output
RC4739	2	40	PNP	3	1	110	± 18	100	D	Raytheon device only Low noise audio amplifier
uA739	1	300	NPN	10	1	86	± 18	90	D	Fairchild device only Low noise audio amplifier Needs frequency compensation
LM381	Not applicable		NPN	15	–	112	± 20	–	D	Low noise amplifier Internally compensated
CA3130	8	0.005	MOSFET	15	10	110	+16	90	S	Ground sensing inputs Very high input impedance Needs frequency compensation

Op amp types	Input offset voltage mV	Input bias current nA	Type of input structure	Band-width MHz	Slew rate V/ns	Voltage gain dB	Maximum supply voltage V	CMRR dB	Qty	COMMENTS
CA3140	8	0.010	MOSFET	4.5	9	100	+36	90	S	Ground sensing inputs Very high input impedance
CA3160	6	0.005	MOSFET	4	10	110	+15	90	S	Ground sensing inputs Very high input impedance
NE531 RC4531	2	400	NPN	10	35	96	± 22	100	S	Very fast op amp Needs frequency compensation
										Pin for pin replacement for
TL080	15	0.4	JFET	3	13	83	± 18	70	S	748
TL081	15	0.4	JFET	3	13	83	± 18	70	S	JFET input op amps, with — 741
TL082	15	0.4	JFET	3	13	83	± 18	70	D	fast slew rate and wide — 1458
TL083	15	0.4	JFET	3	13	83	± 18	70	D	bandwidth [Texas] — 747
TL084	15	0.4	JFET	3	13	83	± 18	70	Q	324









TL080 bifet op-amps

The TL080 family of BIFET operational amplifiers, provides an ideal combination of high-impedance JFET inputs with a low-distortion bipolar output circuit. Quality performance in the TL080 family is achieved without complex circuitry.

TL080 family circuit description

The following sections should be read in conjunction with Fig 1, the basic schematic for one channel.

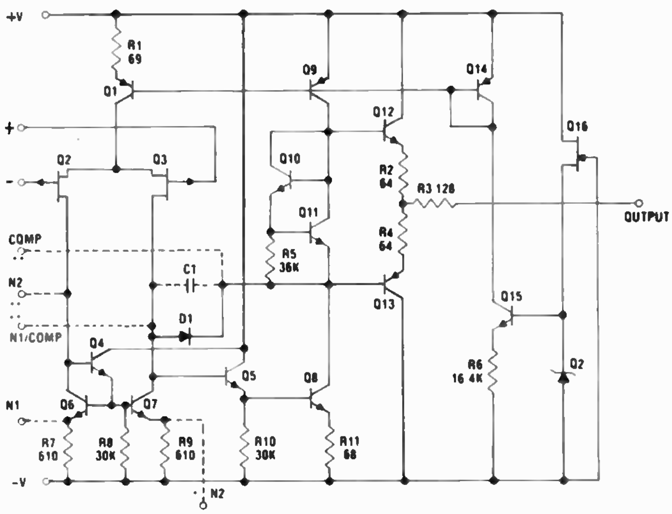


Fig 1 Schematic diagram for TL080 family.

Bias circuits

EFT Q16, zener D2, transistors Q14/Q15 and resistor R6 establish the bias currents for the input differential amplifier and the second gain stage. Epitaxial FET Q16 provides a fixed current to D2 establishing 5.2V on the base of Q15. The resulting 317uA collector current of Q15 flows through Q14 and sets the current levels in Q1 and Q9.

Resistor R1 causes 196uA current in Q1 that is divided between the input stage JFETs Q2 and Q3. The second-gain-stage bias current, about 600uA, is derived from E9.

Second stage

Drive from the input stage is single-ended from the collector of Q7. D1 provides a clamping action across Q5 and Q8 preventing saturation of Q8 and excessive current in Q5. Q5 and Q8 form the high-gain second stage. The second stage output, collector of Q8, drives the output stage consisting of bias transistors Q10 and Q11, and output drivers Q12 and Q13.

Input circuit

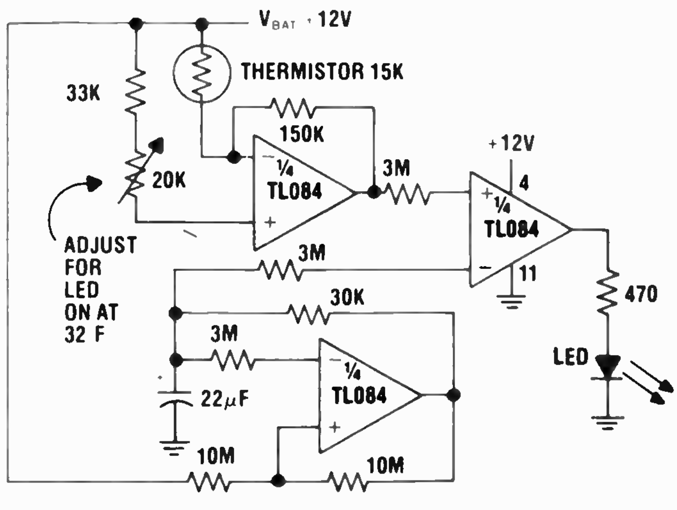
Input JFETs Q2 and Q3 operate into the active load circuit consisting of Q4, Q6, and Q7. Current imbalance and input offset voltages may be adjusted on the TL081 and TL083 through connections to the emitters of Q6 and Q7. External offset controls for the TL080 connect to the collectors of Q6 and Q7. The C1 compensation capacitor is internal on the TL080, TL082 and TL083, and TL084. For the TL080 connections for external compensation are provided which allow user adjustment of AC characteristics.

Ion-implanted input devices provide very high input impedance, controlled pinch-off voltage for maximum common-mode input range, and matched characteristics for control of the input offset voltage. JFET inputs also allow adequate drive to the second stage resulting in maximum output peak-to-peak capability and wide power band widths.

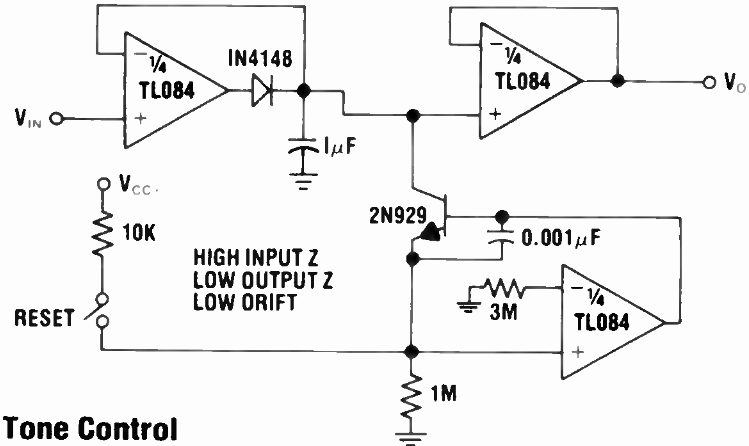
Output stage

Q10 and Q11 provide Class AB bias to the output transistors Q12 and Q13. This allows near zero crossover distortion and produces a low total harmonic distortion at the output. The simplicity of the output circuit results in minimum silicon area requirements keeping manufacturing cost down while maintaining quality performance. R2, R3 and R4 form the output short-circuit protection network.

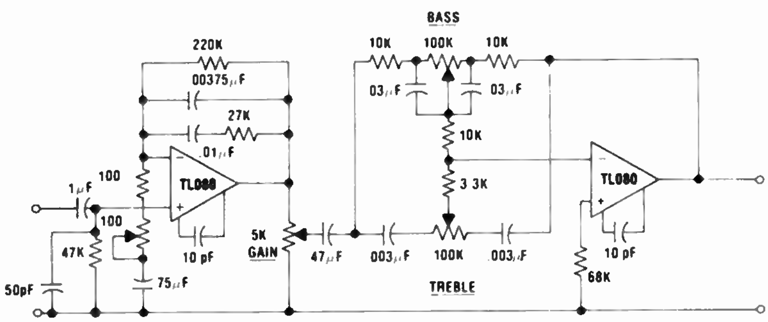
Icy Road Warning Indicator



Peak Detector



Tone Control

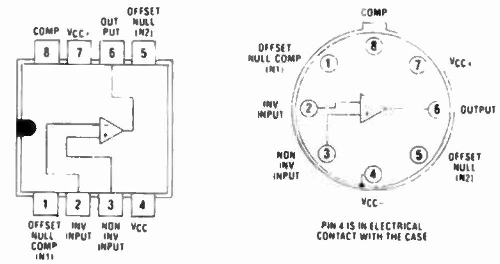


FEATURES	ADVANTAGES
● HIGH INPUT IMPEDANCE	Minimum loading effects allow efficient use with high impedance transducers.
● HIGH SLEW RATE	Provides the desired response characteristics required in audio frequency active filters and quality sound systems.
● LOW DISTORTION	Minimized crossover distortion yields very low total harmonic distortion for maximum performance in critical music systems.
● CONTINUOUS SHORT CIRCUIT PROTECTION	No damage resulting from continual shorts or operation into low impedance loads
● LOW POWER CONSUMPTION	Only 2.8 mA per operational amplifier. Less system power required and battery operation is practicable

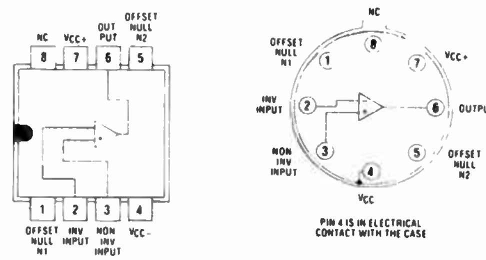
absolute maximum ratings		TL08_C TL08_AC TL08_BC
Supply voltage, V_{CC} (see Note 1)		18 V
Supply voltage, V_{CC} (see Note 1)		−18 V
Differential input voltage (see Note 2)		±30 V
Input voltage (see Notes 1 and 3)		±15 V
Duration of output short circuit (see Note 4)		Unlimited
Continuous total dissipation at 25°C free-air temperature	J,JG,N, or P Package	680 mW
	L Package	625 mW
Operating free-air temperature range		0 to 70°C

PIN OUTS

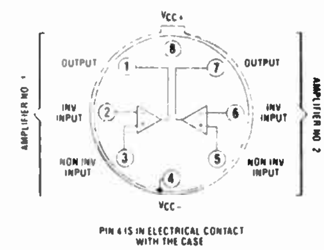
TL080



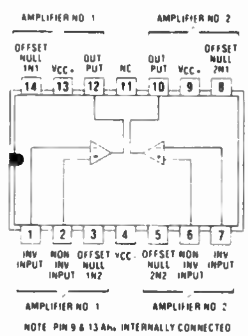
TL081



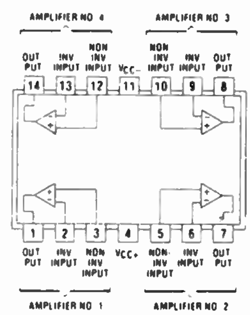
The TL082



The TL083



The TL084



NOTES:
1. All voltage values, except differential voltages, are with respect to the zero reference level (ground) of the supply voltages where the zero reference level is the midpoint between V_{CC+} and V_{CC-}
2. Differential voltages are at the noninverting input terminal with respect to the inverting input terminal.
3. The magnitude of the input voltage must never exceed the magnitude of the supply voltage or 15 volts, whichever is less.
4. The output may be shorted to ground or to either supply. Temperature and/or supply voltages must be limited to ensure that the dissipation rating is not exceeded.

μA741 frequency-compensated operational amplifier

QUICK REFERENCE DATA

Maximum ratings

Supply voltage	± 18 V
Internal power dissipation, metal can	500 mW
Internal power dissipation ceramic DIP	670 mW
Internal power dissipation, silicone DIP	340 mW
Internal power dissipation, mini DIP	310 mW
Internal power dissipation, flatpack	570 mW
Differential input voltage	± 30 V
Input voltage	± 15 V
Operating temperature range	0°C to 70°C
Output short-circuit duration	Indefinite

Characteristics (V = ± 15V)

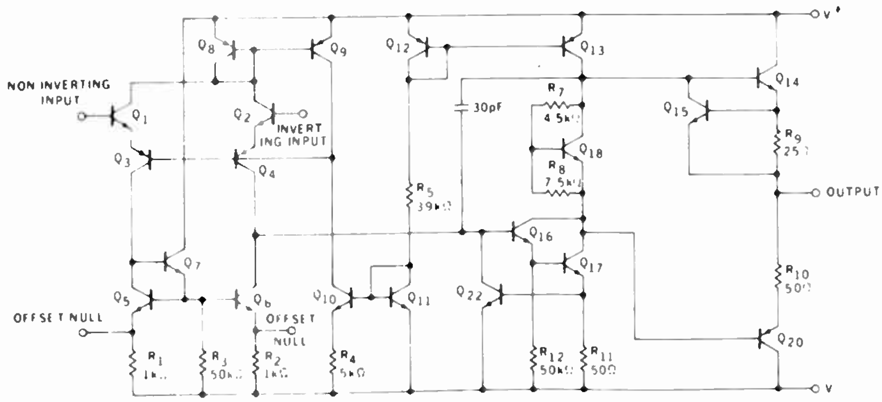
Input offset voltage ($R_S \leqslant 10\ k\Omega$)	typ. 2 mV
Input offset current	typ. 20 nA
Input bias current	typ. 80 nA
Input resistance	typ. 2 MΩ
Input capacitance	typ. 1.4 pF
Offset voltage adjustment range	typ. ± 15 mV
Input voltage range	typ. ± 13 V
Common mode rejection ratio ($R_S \leqslant 10\ k\Omega$)	typ. 90 dB
Supply voltage rejection ratio ($R_S \leqslant 10\ k\Omega$)	typ. 30 µV/V
Large-signal voltage gain ($R_L \geqslant 2\ k\Omega$, $V_{OUT} = \pm 10$ V)	typ. 200 000
Output voltage swing ($R_L \geqslant 10\ k\Omega$)	typ. ± 14 V
Output voltage swing ($R_L \geqslant 2\ k\Omega$)	typ. ± 13 V
Output resistance	typ. 75 Ω
Output short circuit current	typ. 25 mA
Supply current	typ. 1.7 mA
Power consumption	typ. 50 mW
Transient response, unity gain (V_{in} = 20 mV, R_L = 2 kΩ, $C_L \leqslant 100$ pF)	
risetime	typ. 0.3 µs
overshoot	typ. 5%
Slew rate ($R_L \geqslant 2\ k\Omega$)	typ. 0.5 V/µs
Input offset voltage	max. 7.5 mV
Input offset current	max. 300 nA
Input bias current	max. 800 nA
Large-signal voltage gain ($R_L \geqslant 2\ k\Omega$, $V_{out} = \pm 10$ V)	min. 15 000
Output voltage swing ($R_L \geqslant 2\ k\Omega$)	typ. ± 13 V

The µA741 is a high performance monolithic operational amplifier. It is suitable for a wide range of analogue applications. As a voltage follower it is ideal; the common mode voltage range is wide and there is no latch-up. The high gain and wide range of operating voltages provides superior performance in integrator, summing amplifier, and general feedback applications.

The 741 requires no frequency compensation and is internally protected against short-circuits. It is by far the most common op-amp in amateur use.

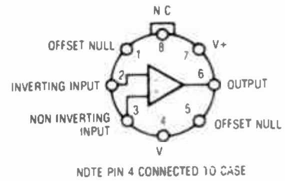
The device comes in two grades — military (312) and commercial (393). Here we deal only with the commercial grade.

EQUIVALENT CIRCUIT

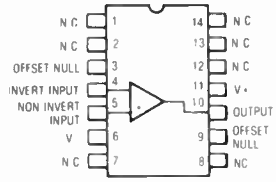


CONNECTION DIAGRAMS
(TOP VIEW)

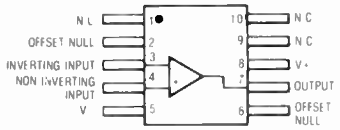
8 LEAD METAL CAN



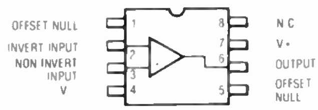
14 LEAD DIP



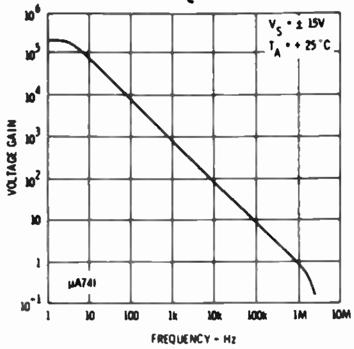
FLATPACK



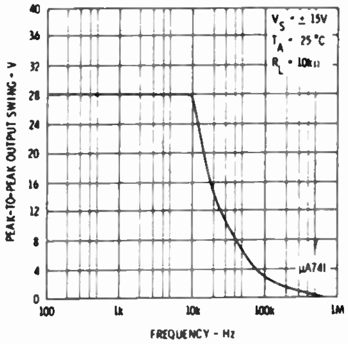
MINIDIP



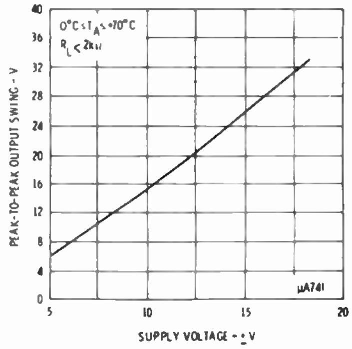
OPEN LOOP VOLTAGE GAIN AS A FUNCTION OF FREQUENCY



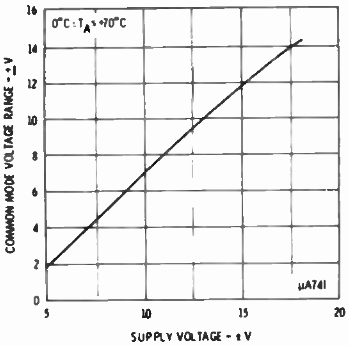
OUTPUT VOLTAGE SWING AS A FUNCTION OF FREQUENCY



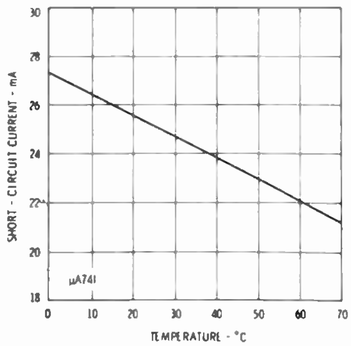
OUTPUT VOLTAGE SWING AS A FUNCTION OF SUPPLY VOLTAGE



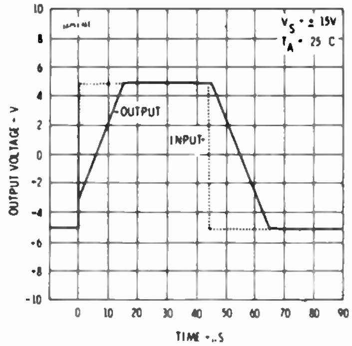
INPUT COMMON MODE VOLTAGE RANGE AS A FUNCTION OF SUPPLY VOLTAGE



OUTPUT SHORT-CIRCUIT CURRENT AS A FUNCTION OF AMBIENT TEMPERATURE

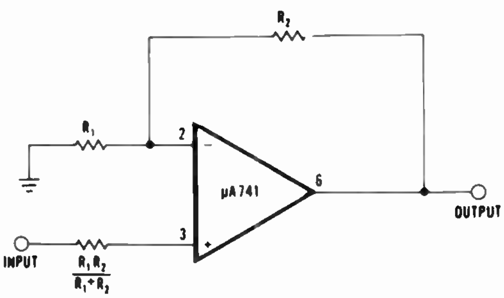


VOLTAGE FOLLOWER LARGE-SIGNAL PULSE RESPONSE



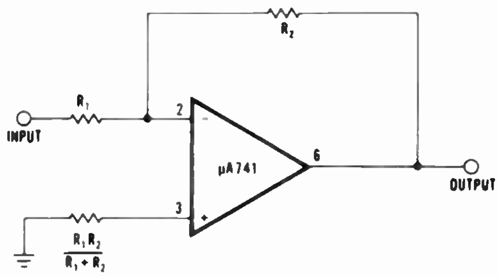
TYPICAL APPLICATIONS

Non-inverting Amplifier

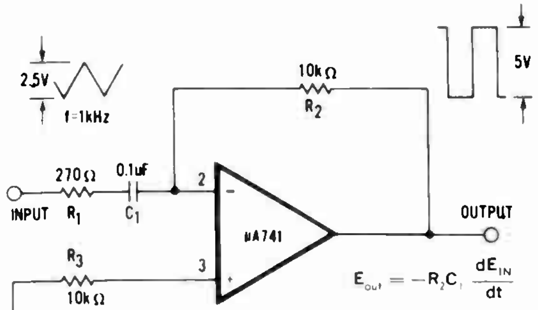


GAIN	R_1	R_2	B.W.	R_{IN}
10	1 kΩ	9 kΩ	100 kHz	400 MΩ
100	100 Ω	9.9 kΩ	10 kHz	280 MΩ
1000	100 Ω	99.9 kΩ	1 kHz	80 MΩ

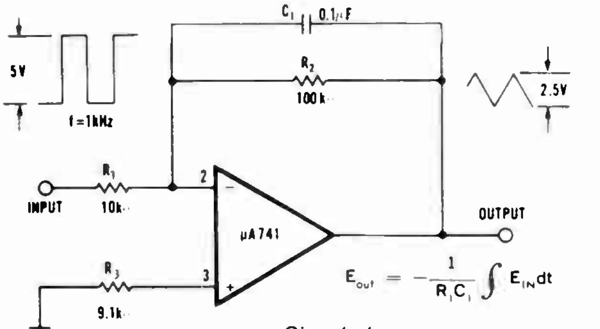
Inverting Amplifier



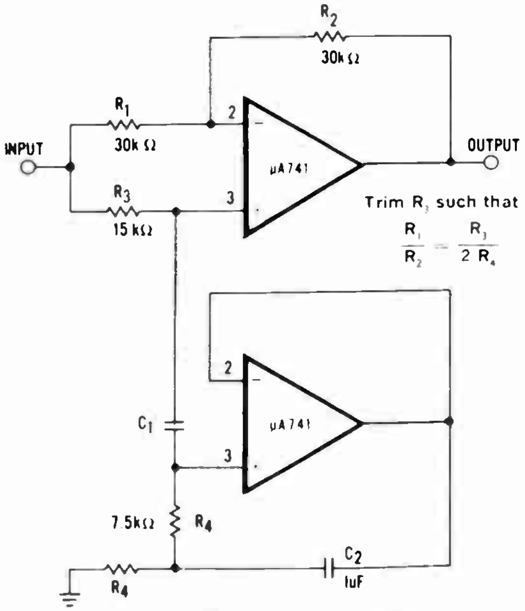
GAIN	R_1	R_2	B.W.	R_{IN}
1	10 kΩ	10 kΩ	1 MHz	10 kΩ
10	1 kΩ	10 kΩ	100 kHz	1 kΩ
100	1 kΩ	100 kΩ	10 kHz	1 kΩ
1000	100 Ω	100 kΩ	1 kHz	100 Ω



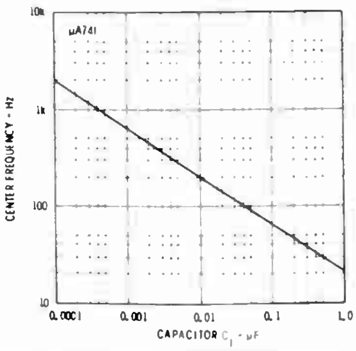
Simple Differentiator

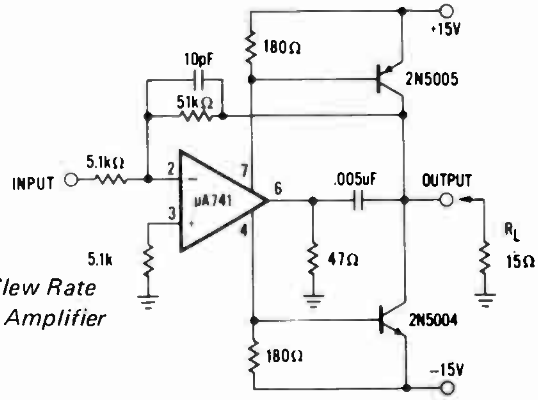


Simple Integrator

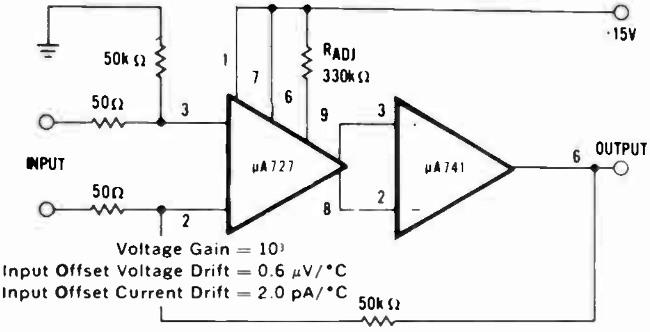


Notch Filter Using the 741 As A Gyrator





High Slew Rate Power Amplifier

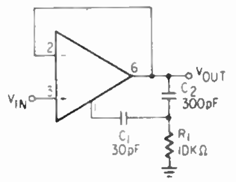


Low Drift Low Noise Amplifier

TYPICAL APPLICATIONS

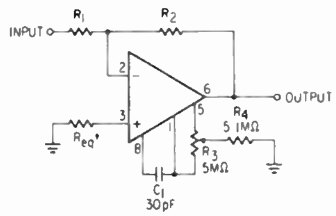
(All pin numbers shown refer to 8-lead TO-5 package)

FAST VOLTAGE FOLLOWER



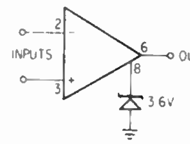
Power Bandwidth: 15 kHz
Slew Rate: 1 V/μs

INVERTING AMPLIFIER WITH BALACING CIRCUIT

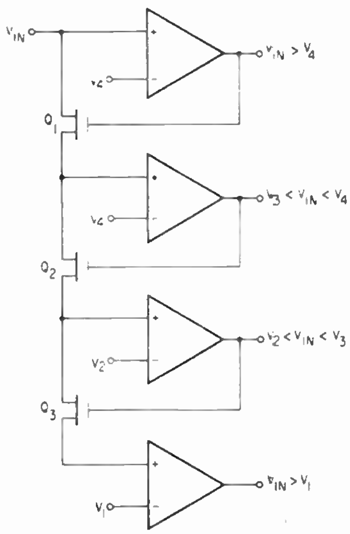


† May be zero or equal to parallel combination of R1 and R2 for minimum offset.

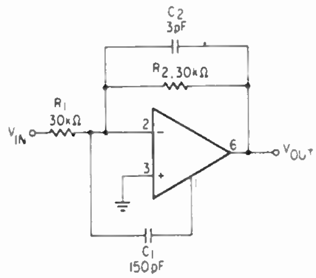
VOLTAGE COMPARATOR FOR DRIVING DTL OR TTL INTEGRATED CIRCUITS



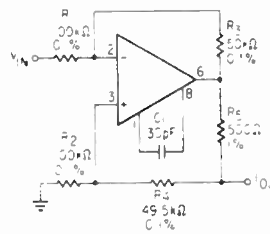
MULTIPLE APERTURE WINDOW DISCRIMINATOR



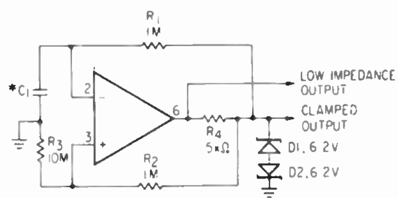
FAST SUMMING AMPLIFIER



BILATERAL CURRENT SOURCE

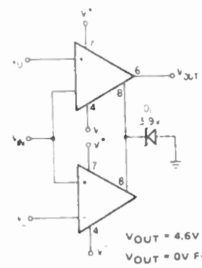


LOW FREQUENCY SQUARE WAVE GENERATOR



* Adjust C_1 for frequency

DOUBLE ENDED LIMIT DETECTOR



V_{OUT} = 4.6V FOR $V_L \leq V_{IN} \leq V_U$
V_{OUT} = 0V FOR $V_{IN} < V_L$ OR $V_{IN} < V_U$

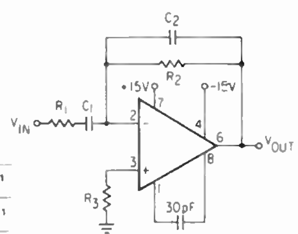
PRACTICAL DIFFERENTIATOR

$f_c = \dfrac{1}{2\pi R_2 C_1}$

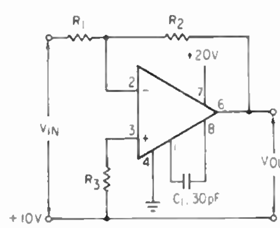
$f_h = \dfrac{1}{2\pi R_1 C_1}$

$\dfrac{1}{2\pi R_2 C_2}$

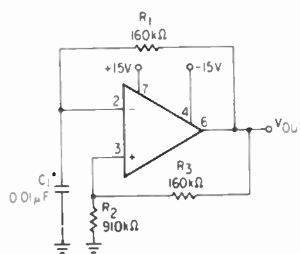
$f_c < f_h < f_{unity\ gain}$



CIRCUIT FOR OPERATING WITHOUT A NEGATIVE SUPPLY

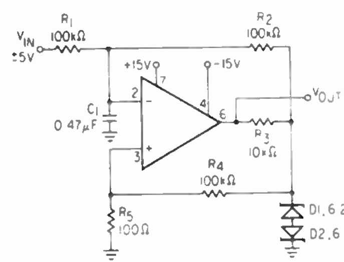


FREE-RUNNING MULTIVIBRATOR



*Chosen for oscillation at 100 Hz

PULSE WIDTH MODULATOR



μA301 general purpose op-amp.

QUICK REFERENCE GUIDE

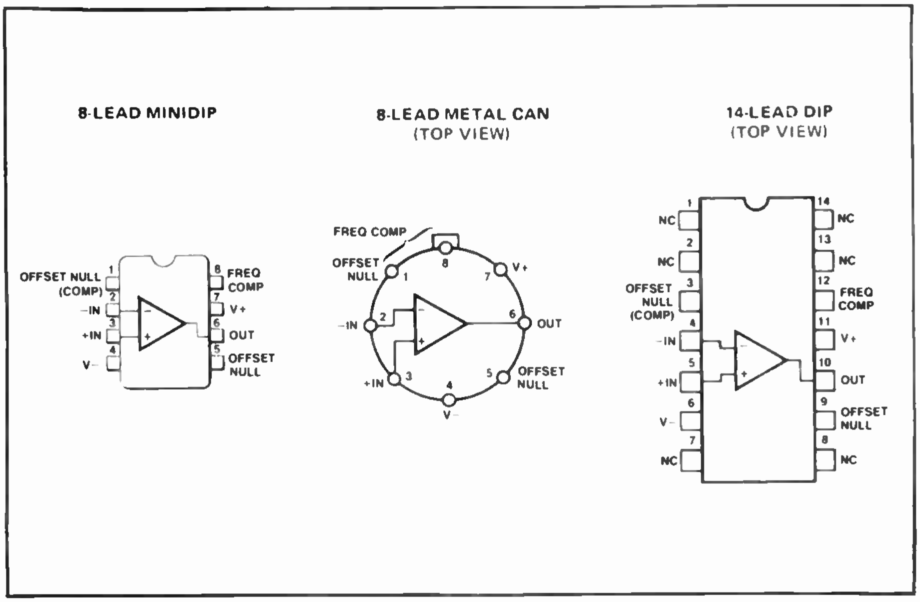
Maximum ratings

Supply voltage	max. 18 V
Internal power dissipation	
Metal can	max. 500 mW
DIP	max. 670 mW
Mini DIP	max. 310 mW
Differential input voltage	max. ±30 V
Input voltage	max. ±15 V
Output short-circuit duration	indefinite

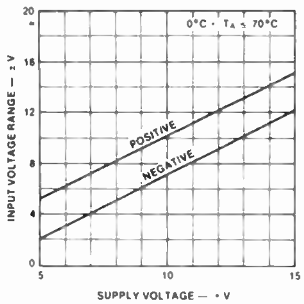
Characteristics
(V_S = 5 to 15 V; C1 = 3p)

Input offset voltage ($R_S \leqslant 10$ k)	typ. 2 mV
Input offset current	typ. 3 nA
Input bias current	typ. 70 nA
Input resistance	typ. 2 M
Supply current (V_S = 15 V)	typ. 1.8 mA
Large signal voltage gain ($V_S = \pm 15$ V; $V_{OUT} = \pm 10$ V; $R_L \geqslant 2$ k)	typ. 160 V/mV
Output voltage swing	
$V_S = \pm 15$ V; R_L = 10 k	typ. ±14 V
V_S = 15 V; R_L = 2 k	typ. ±13 V
Input voltage range ($V_S = \pm 15$ V)	typ. ±12 V
Common mode rejection ratio ($R_S \leqslant 10$ k)	typ. 90 dB
Supply voltage rejection ratio ($R_S \leqslant 10$ k)	typ. 90 dB

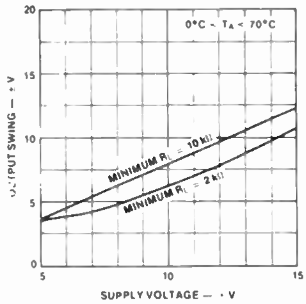
These op-amps (LM301, μA301, etc are intended for applications requirin low input offset voltage or low inpu offset current. The low drift and bia currents are advantageous in suc applications as long interval integrator timers and sample and hold circuits. Th frequency response can be set with on external capacitor. There is no latch-u and protection against short circuit



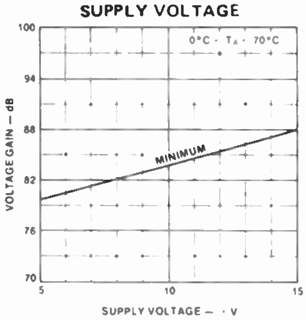
INPUT VOLTAGE RANGE AS A FUNCTION OF SUPPLY VOLTAGE



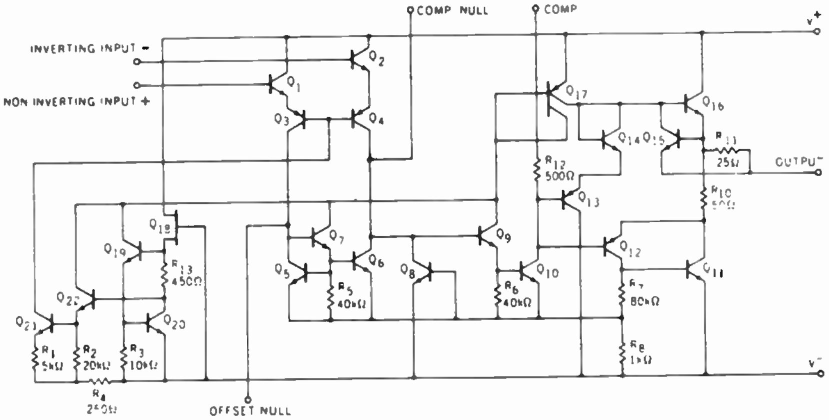
OUTPUT SWING AS A FUNCTION OF SUPPLY VOLTAGE



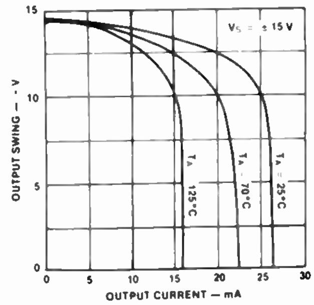
VOLTAGE GAIN AS A FUNCTION OF SUPPLY VOLTAGE



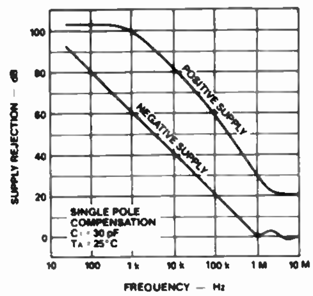
EQUIVALENT CIRCUIT



Current Limiting

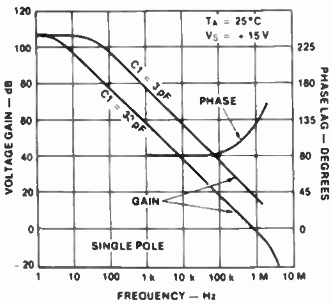


POWER SUPPLY REJECTION AS A FUNCTION OF FREQUENCY

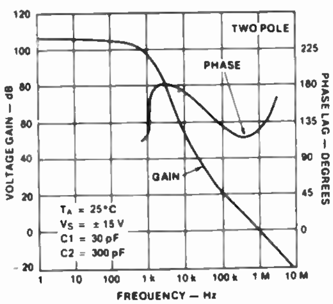


TYPICAL PERFORMANCE CURVES

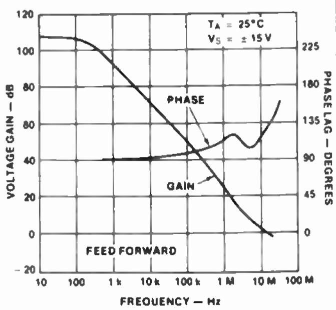
OPEN LOOP FREQUENCY RESPONSE



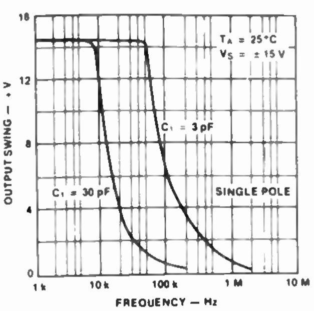
OPEN LOOP FREQUENCY RESPONSE



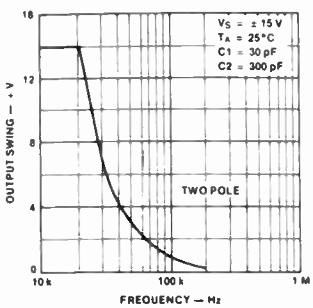
OPEN LOOP FREQUENCY RESPONSE



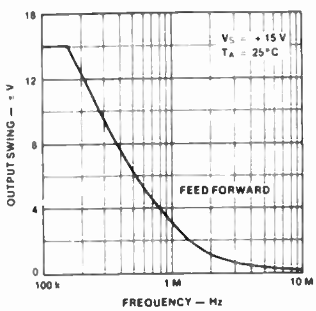
LARGE SIGNAL FREQUENCY RESPONSE



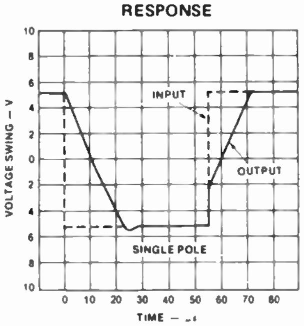
LARGE SIGNAL FREQUENCY RESPONSE



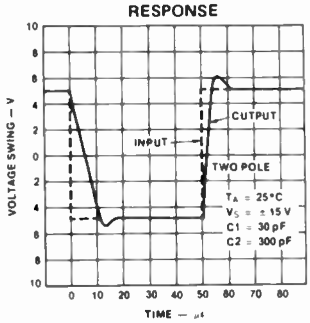
LARGE SIGNAL FREQUENCY RESPONSE



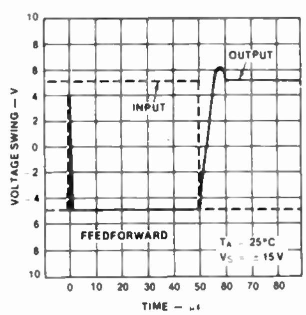
VOLTAGE FOLLOWER PULSE RESPONSE



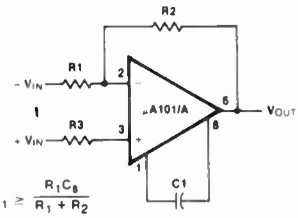
VOLTAGE FOLLOWER PULSE RESPONSE



INVERTER PULSE RESPONSE



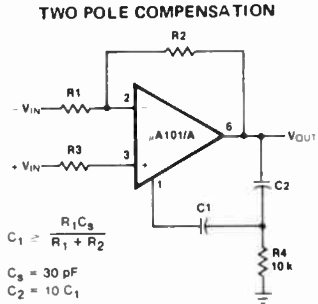
SINGLE POLE COMPENSATION



$C_1 \geq \frac{R_1 C_S}{R_1 + R_2}$

$C_S = 30$ pF

TWO POLE COMPENSATION

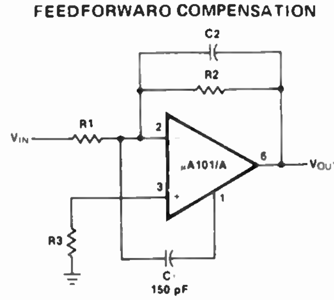


$C_1 \geq \frac{R_1 C_S}{R_1 + R_2}$

$C_S = 30$ pF

$C_2 = 10\ C_1$

FEEDFORWARD COMPENSATION



NATIONAL LM324

The LM124 series op amps operate with only a single power supply and have true-differential inputs. They remain in the linear mode with an input common-mode voltage of 0 V_{DC}.

Precautions should be taken to insure that the power supply for the integrated circuit never becomes reversed in polarity.

Large differential inputs voltages can b easily accomodated and, as input di ferential voltage protection diodes ar not needed, no large input current result from large differential input vo tages. The differential input voltage ma be larger than V^+ without damagin the device. Protection should be provide to prevent the input voltages from goin negative more than $-0.3\,V_{DC}$ (at 25°C An input clamp diode with a resistor t the IC input terminal can be used.

To reduce the power supply curren drain, the amplifiers have a class A ou put stage for small signal levels whic converts to class B in a large signal mode This allows the amplifiers to bot source and sink large output current Therefore both NPN and PNP externa current boost transistors can be used t extend the power capability of the basi amplifiers.

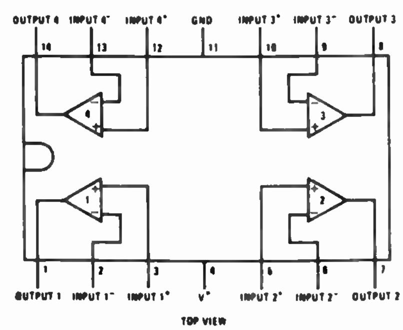
For ac applications, where the load i

electrical characteristics ($V^+ = +5.0\,V_{DC}$

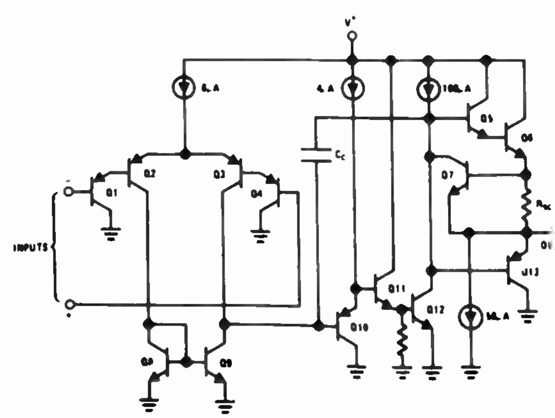
PARAMETER	CONDITIONS	LM324 MIN	TYP	MAX	UNITS
Input Offset Voltage	$T_A = 25°C$		±2	±7	mV_{DC}
Input Bias Current	$I_{IN(+)}$ or $I_{IN(-)}$, $T_A = 25°C$		45	250	nA_{DC}
Input Offset Current	$I_{IN(+)} - I_{IN(-)}$, $T_A = 25°C$		±5	±50	nA_{DC}
Input Common-Mode Voltage Range	$V^+ = 30\,V_{DC}$, $T_A = 25°C$	0		$V^+-1.5$	V_{DC}
Supply Current	$R_L = \infty$, $V_{CC} = 30V$, (LM2902 $V_{CC} = 26V$) $R_L = \infty$ On All Op Amps Over Full Temperature Range $T_A = 25°C$		1.5 0.7	3 1.2	mA_{DC} mA_{DC} mA_{DC}
Large Signal Voltage Gain	$V^+ = 15\,V_{DC}$ (For Large V_O Swing) $R_L \geq 2\,k\Omega$, $T_A = 25°C$	25	100		V/mV
Output Voltage Swing	$R_L = 2\,k\Omega$, $T_A = 25°C$ (LM2902 $R_L \geq 10\,k\Omega$)	0		$V^+-1.5$	V_{DC}
Common-Mode Rejection Ratio	DC, $T_A = 25°C$	65	70		dB
Power Supply Rejection Ratio	DC, $T_A = 25°C$	65	100		dB
Amplifier-to-Amplifier Coupling	f = 1 kHz to 20 kHz, $T_A = 25°C$ (Input Referred)		−120		dB
Output Current Source	$V_{IN}^+ = 1\,V_{DC}$, $V_{IN}^- = 0\,V_{DC}$, $V^+ = 15\,V_{DC}$, $T_A = 25°C$	20	40		mA_{DC}
Sink	$V_{IN}^- = 1\,V_{DC}$, $V_{IN} = 0\,V_{DC}$, $V^+ = 15\,V_{DC}$, $T_A = 25°C$	10	20		mA_{DC}
	$V_{IN}^- = 1\,V_{DC}$, $V_{IN}^+ = 0\,V_{DC}$, $T_A = 25°C$, $V_O = 200\,mV_{DC}$	12	50		μA_{DC}
Short Circuit to Ground	$T_A = 25°C$		40	60	mA_{DC}
Input Offset Voltage				±9	mV_{DC}
Input Offset Voltage Drift	$R_S = 0\Omega$		7		$\mu V/°C$
Input Offset Current	$I_{IN(+)} - I_{IN(-)}$			±150	nA_{DC}
Input Offset Current Drift			10		$pA_{DC}/°C$
Input Bias Current	$I_{IN(+)}$ or $I_{IN(-)}$		40	500	nA_{DC}
Input Common-Mode Voltage Range	$V^+ = 30\,V_{DC}$	0		V^+-2	V_{DC}
Large Signal Voltage Gain	$V^+ = +15\,V_{DC}$ (For Large V_O Swing) $R_L \geq 2\,k\Omega$	15			V/mV
Output Voltage Swing V_{OH}	$V^+ = +30\,V_{DC}$, $R_L = 2\,k\Omega$ $R_L \geq 10\,k\Omega$	26 27	 28		V_{DC} V_{DC}
V_{OL}	$V^+ = 5\,V_{DC}$, $R_L \leq 10\,k\Omega$		5	20	mV_{DC}
Output Current Source	$V_{IN}^+ = +1\,V_{DC}$, $V_{IN}^- = 0\,V_{DC}$, $V^+ = 15\,V_{DC}$	10	20		mA
Sink	$V_{IN}^- = +1\,V_{DC}$, $V_{IN}^+ = 0\,V_{DC}$, $V^+ = 15\,V_{DC}$	5	8		mA
Differential Input Voltage				V^+	V_{DC}

connection diagram

Dual-In-Line and Flat Package



schematic diagram (Each Amplifier)



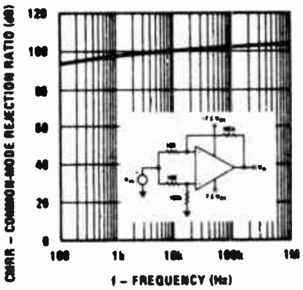
capacitively coupled to the output of the amplifier, a resistor should be used, from the output of the amplifier to ground to increase the class A bias current and prevent crossover distortion. Where the load is directly coupled, as in dc applications, there is no crossover distortion.

Large closed loop gains or resistive isolation should be used if load capacitances over 50 pF must be driven by the amplifier.

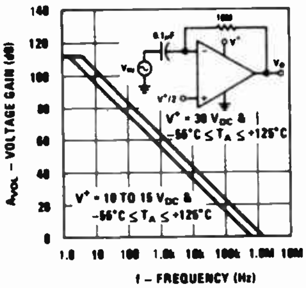
Output short circuits either to ground or to the positive power supply should be of short time duration. Putting direct short-circuits on more than one amplifier at a time will increase the total IC power dissipation to destructive levels, if not properly protected with external dissipation limiting resistors in series with the output leads of the amplifiers.

Introducing a pseudo-ground (a bias voltage reference of $V^+/2$) will allow operation above and below this value in single power supply systems.

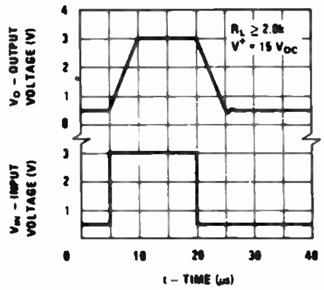
Common Mode Rejection Ratio



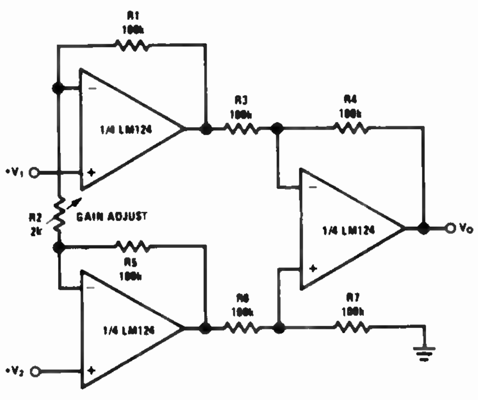
Open Loop Frequency Response



Voltage Follower Pulse Response



High Input Z Adjustable-Gain DC Instrumentation Amplifier

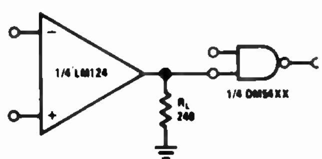


If R1 = R5 & R3 = R4 = R6 = R7 (CMRR depends on match)

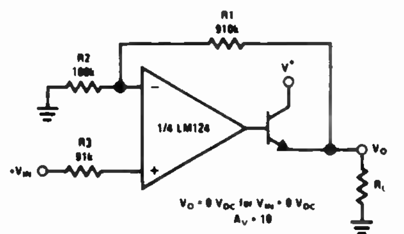
$$V_O = 1 + \frac{2R1}{R2}(V_2 - V_1)$$

As shown $V_O = 101\ (V_2 - V_1)$

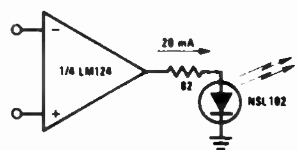
Driving TTL



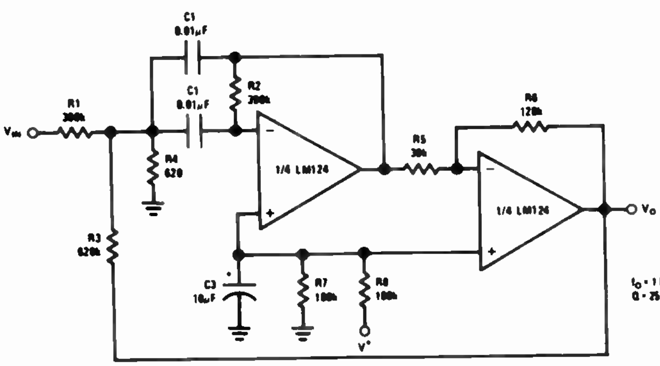
Power Amplifier



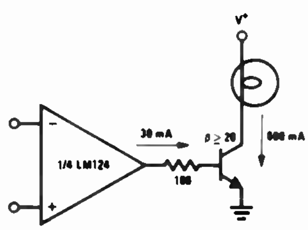
LED Driver



Bandpass Active Filter



Lamp Driver



CA 3140 operational amplifier

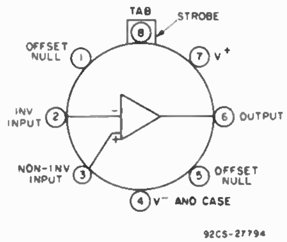


Fig.1 – Functional diagram of CA3140 series.

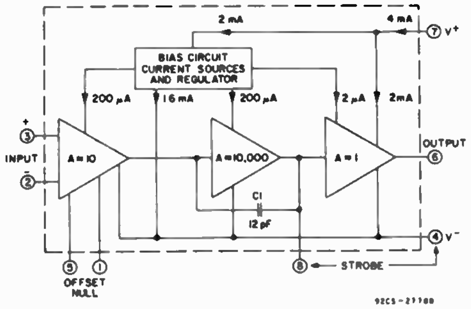


Fig.2 – Block diagram of CA3140 series.

TYPICAL ELECTRICAL CHARACTERISTICS

CHARACTERISTIC		TEST CONDITIONS $V^+ = +15$ V $V^- = -15$ V $T_A = 25^oC$		UNITS
Input Resistance	R1		1.5	TΩ
Input Capacitance	C1		4	pF
Output Resistance	RO		60	Ω
Equivalent Wideband Input Noise Voltage	en	BW = 140 kHz R_S = 1 MΩ	48	μV
Equivalent Input Noise Voltage	e_n	f = 1 kHz R_S =	40	
		f = 10 kHz 100Ω	12	
Short-Circuit Current to Opposite Supply Source	10M+		40	mA
Sink	10M−		18	mA
Gain-Bandwidth Product 8	f_T		4.5	MHz
Slew Rate	SR		9	V/μs
Sink Current From Terminal 8 To Terminal 4 to Swing Output Low	8		220	μA
Transient Response: Rise Time 8	t_r	R_L = 2kΩ	0.08	μs
Overshoot		C_L = 100 pF	10	%
Settling Time at 10 Vpp 1 mV	t_s	R_L = 2kΩ	4.5	μs
10 mV		C_L = 100 pF Voltage Follower	1.4	

MAXIMUM RATINGS, Absolute-Maximum Values.

DC SUPPLY VOLTAGE
(BETWEEN V+ AND V- TERMINALS) . . . . .36 V
DIFFERENTIAL-MODE INPUT VOLTAGE . . . . ±8 V
COMMON-MODE DC INPUT VOLTAGE . . . . (V+ +8V) to (V- -0.5V)
INPUT-TERMINAL CURRENT . . . . 1 mA
DEVICE DISSIPATION:
WITHOUT HEAT SINK –
UP TO 55°C. . . . . .630 mW
ABOVE 55°C . . . . Derate linearly 6.67 mW/°C
WITH HEAT SINK –
Up to 55°C . . . . 1 W
Above 55°C . . . . Derate linearly 16.7 mW/°C
OUTPUT SHORT-CIRCUIT DURATION .INDEFINITE

Short circuit may be applied to ground or to either supply.

The CA3140B, CA3140A, and CA3140 ar integrated-circuit operational amplifiers tha combine the advantages of high-voltage PMO transistors with high-voltage bipolar trar sistors on a single monolithic chip. Becaus of this unique combination of technologie this device can now provide designers, fc the first time, with the special performanc features of the CA3130 COS/MOS oper tional amplifiers and the versatility of th 741 series of industry-standard operation amplifiers.

The CA3140, CA3140A, and CA3140 BiMO operational amplifiers feature gate-protecte MOS/FET (PMOS) transistors in the inpu circuit to provide very-high-input impedanc very-low-input current, and high-speed pe formance. The CA3140B operates at suppl voltages from 4 to 44 volts; the CA3140 and CA3140 from 4 to 36 volts (either singl or dual supply). These operational amplifier are internally phase-compensated to achiev stable operation in unity-gain follower oper ation, and, additionally, have access termi nals for a supplementary external capacito if additional frequency roll-off is desired Terminals are also provided for use in appli cations requiring input offset-voltage nulling

The use of PMOS field-effect transistors i the input stage results in common-mode in put-voltage capability down to 0.5 volt belov the negative-supply terminal, an importan attribute for single-supply applications. Th output stage uses bipolar transistors and in cludes built-in protection against damag from load-terminal short-circuiting to eithe supply-rail or to ground.

The CA3140 Series has the same 8-lead termi nal pin-out used for the "741" and othe industry-standard operational amplifiers.The are supplied in either the standard 8-lea TO-5 style package (T suffix), or in the 8 lead dual-in-line formed-lead TO-5 style pack age "DIL-CAN" (S suffix). The CA3140E is intended for operation at supply volt ages ranging from 4 to 44 volts, for appli cations requiring premium-grade specifica tions and with electrical limits establishe for operation over the range from −55°C to + 125°C. The CA3140A and CA314 are for operation at supply voltages up t 36 volts (±18 volts). The CA3140 an CA3140A can also be operated safely ove the temperature range from −55°C to 125°C, although specification limits for thei electrical parameters do not apply whe they are operated beyond their specifie temperature ranges.

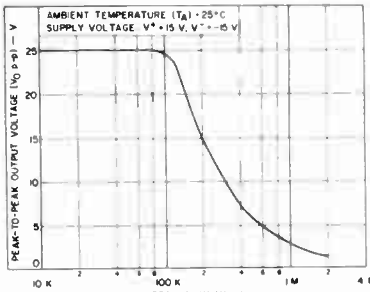


Fig.8 — Maximum output voltage swing vs frequency.

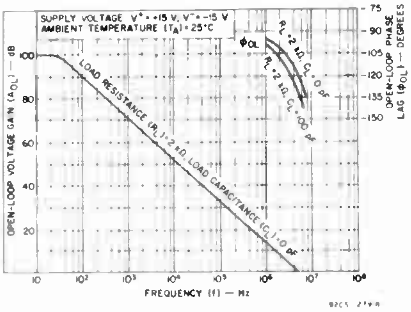


Fig. 18 — Open-loop voltage gain and phase lag vs frequency.

ELECTRICAL CHARACTERISTICS FOR EQUIPMENT DESIGN

At V^+ = 15 V, V^- = 15 V, T_A = 25°C Unless Otherwise Specified

CHARACTERISTIC	Min.	Typ.	Max.	UNITS
Input Offset Voltage, V10	–	5	15	mV
Input Offset Current I_{10}	–	0.5	30	pA
Input Current I_I	–	10	50	pA
Large-Signal Voltage Gain, A_{OL}*	86	100	–	dB
Common-Mode Rejection Ratio, CMRR	70	90	–	dB
Common-Mode Input-Voltage Range, V_{ICR}	–15	–15.5 to +12.5	11	V
At V^+ = 5 V, V^- = 0 V,	–0.5 to +2.6			V
Power Supply Rejection Ratio, PSRR (see Fig. 11)	76	80	–	dB
Max. Output Voltage■	V_{OM+}	+12	13	–
	V_{OM-}	–14	–14.4	–
Supply Current, I^t	–	4	6	mA
Device Dissipation, P_D	–	120	180	mW

★ At V_O = 26$V_{p\text{-}p}$, +12V, –14V and R_L = 2kΩ

■ At R_L = 2kΩ

INPUT CIRCUIT CONSIDERATIONS

As mentioned previously, the amplifier inputs can be driven below the terminal 4 potential, but a series current-limiting resistor is recommended to limit the maximum input terminal current to less than 1 mA to prevent damage to the input protection circuitry.

Moreover, some current-limiting resistance should be provided between the inverting input and the output when the CA3140 is used as a unity-gain voltage follower. This resistance prevents the possibility of extremely large input-signal transients from forcing a signal through the input-protection network and directly driving the internal constant-current source which could result in positive feedback via the output terminal. A 3.9-kΩ resistor is sufficient.

The typical input current is in the order of 10 pA when the inputs are centered at nominal device dissipation. As the output supplies load current, device dissipation will increase, raising the chip temperature and resulting in increased input current.

DRIVING TTL

Excellent interfacing with TTL circuitry is easily achieved with a single 6.2-volt zener diode connected to terminal 8 as shown in Fig 5. This connection assures that the maximum output signal swing will not go more positive than the zener voltage minus two base-to-emitter voltage drops within the CA3140.

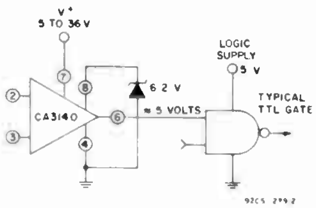


Fig.5. Zener clamping diode connected to terminals 8 and 4 to limit the CA3140 output swing to TTL levels.

OFFSET-VOLTAGE NULLING

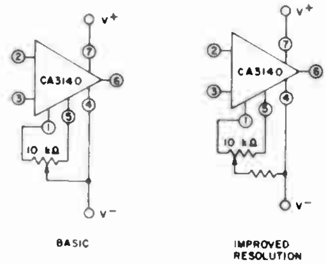


Fig.6. Two offset-voltage nulling methods.

Three-terminal positive voltage regulators

This month's data sheets cover the more common positive three-terminal voltage regulators. The table immediately below is a summary of the information on the other tables. At the end of the data sheets we include notes on heatsinking and a warning on possible problems.

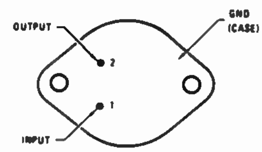
Data sheet summary — Three terminal voltage regulators

Device	V_{OUT} (V)	Line regulation (mV/V)	Load regulation (mV/V)	Max. V_{IN} (V)	Max. I_Q (mA)	Dropout Voltage (V)
LM109, LM209 LM309	5	10	20	35	10	1-2
LM340	5,6,8,12,15,18	20	20	35	10	1.6-2
LM78XX	24	20	20	40	10	1.6-2
LM78LXXAC	5	20	12	30	6	1.5-2
	8,12,15,18	20	12	35	6	1.5-2
	24	20	12	40	6	1.5-2
LM3910	5,6,8,10 12,15,18,24	13	5	40	4	1.5-2

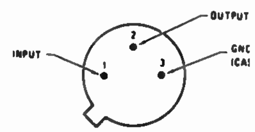
Device	Package	Rated I_{OUT} (A)	Typ. θ_{JC} (°C/W)	Typ. θ_{JA} (°C/W)	Max. P_O (W)
LM109H series	TO39	0.2	15	150	2
LM109K series	TO3	1	3	35	20
LM340K series	TO3	1	4	35	18
LM340T series	TO220	1	4	50	18
LM78LXXACH	TO39	0.1	40	140	3
LM78LXXACZ	TO92	0.1	40	180	1
LM3910H	TO39	0.1	40	140	3
LM3910Z	TO92	0.1		180	1

LM109, 209, 309 five-volt regulators

TO-3 (K)



TO-39 (H)



LM109, LM209, LM309

Junction temperature	LM109	T_J	max. 150°C min. −55°C
	LM209	T_J	max. 150°C min. −25°C
	LM309	T_J	max. 125°C min. +50°C
Thermal resistance	TO3 TO5		typ. 35 C/W typ. 150 °C/W
Thermal resistance, junction-to-case	TO3 TO5	θ_{JC} θ_{JC}	typ. 15 °C/W typ. 3 °C/W
Power dissipation (internally limited)	TO3 TO5	P_{MAX} P_{MAX}	max. 2 W max. 20 W
Output current	TO3	I_O	max. 1 A typ. 0.5 A
	TO5	I_O	max. 0.2 A typ. 0.1 A
Input voltage		V_I	max. 35 V typ. 10 V
Output voltage (V_{IN} = 7 to 25 V; I_O = 5 mA to I_{MAX})		V_O	typ. 5.05 V
Line regulation		ΔV_O	typ. 4 mV
Load regulation	TO3 (I_O = 5 to 1500 mA) TO5 (I_O = 5 to 500 mA)		typ. 50 mV typ. 20 mV
Quiescent current		I_Q	typ. 40 µV
Output noise voltage		vn	typ. 40µV

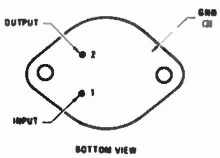
These voltage regulators are complete 5V regulators fabricated on a single silicon chip. They are designed for local regulation on digital logic cards, eliminating the distribution problems associated with single-point regulation. The devices are available in two common transistor packages. In TO-5 it can deliver output currents in excess of 200 mA, if adequate heat sinking is provided. With the TO-3 power package the available output current is greater than 1A.

The regulators are essentially blow out proof. Current limiting is included to limit the peak output current to a safe value in addition, thermal shut down is provided to keep the IC from overheating. If internal dissipation becomes too great, the regulator will shut down to prevent excessive heating.

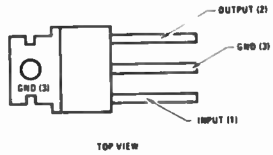
It is not necessary to bypass the output, although this does improve transient response somewhat. Input bypassing is needed, however, if the regulator is located very far from the filter capacitor of the power supply. Stability is also achieved by methods that provide very good rejection of load or line transients as are usually seen with TTL logic.

LM340 & LM78XX series

TO-3 (K)



TO-220 (T)



LM340 -5, -6, -8, -12, -15, -18, -24
LM7805, LM7806, LM7808, LM7812, LM7815, LM7818, LM7824

Output voltage	V_O	typ.	5	6	8	12	15	18	24	V
Input voltage	V_{IN}	typ.	10	11	14	19	23	27	33	V
(for V_O ±4%)		max.	20	21	23	27	30	33	38	V
		min.	7	8	10.5	14.5	17.5	21	27	V
Output current	I_O	typ.	500	500	500	500	500	500	500	mA
Line regulation (I_O = 100 mA)	ΔV_O	max.	50	60	80	120	150	180	240	mV
(I_O = 500 mA)	ΔV_O	max.	100	120	160	240	300	360	480	mV
Load regulation (I_O = 5 to 1500 mA)		max.	100	120	160	240	300	360	480	mV
Quiescent current	I_Q	typ.	7	7	7	7	7	7	7	mA
Output noise voltage	v_n	typ.	40	45	52	75	90	110	170	µV
Ripple rejection (120 Hz)	$\frac{\Delta V_{IN}}{\Delta V_{OUT}}$	typ.	60	57	55	52	50	48	44	dB
Dropout voltage (I_O = 1 A)	$V_{IN}-V_{OUT}$	typ.	2	2	2	2	2	2	2	V

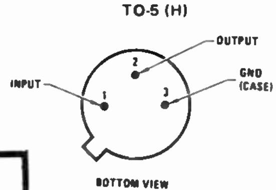
The LM340-XX and LM78XX series of three terminal regulators are available with several fixed output voltages.

The LM340 is available in two power packages. Both the plastic TO-220 and metal TO-3 packages allow these regulators to deliver over 1,0A if adequate heat sinking is provided. Current limiting is included to limit the peak output current to a safe value. Safe area protection for the output transistor is provided to limit internal power dissipation. If internal power dissipation becomes too high for the heat sinking provided, the thermal shutdown circuit takes over preventing the IC from overheating.

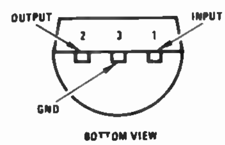
Input bypassing is needed only if the regulator is located far from the filter capacitor of the power supply.

LM78LXX series

TO-5 (H)



TO-92 (Z)



LM78L05, LM78L08, LM78L12, LM78L15, LM78L18, LM78L24

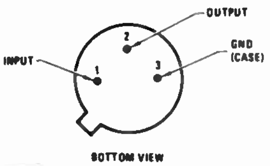
Output voltage	V_O		5	8	12	15	18	24	V
Input voltage	V_{IN}	typ.	10	14	19	23	27	33	V
		max.	30	30	35	35	35	40	V
		min.	7	10.5	14.5	17.5	21.5	27.5	V
Output current	I_O	typ.	40	40	40	40	40	40	mA
Junction temperature	T_J	max.	150	150	150	150	150	150	°C
Thermal resistance	TO5	typ.	140	140	140	140	140	140	°C/W
	TO92	typ.	180	180	180	180	180	180	°C/W
" " junction to case	TO5	typ.	40	40	40	40	40	40	°C/W
Line regulation (I_O low)		typ.	10	12	20	25	27	30	mV
(typ. I_O)		typ.	18	20	30	30	32	35	mV
Load regulation ($I_O \leq$ 40 mA)		typ.	5	6	10	12	15	20	mV
($I_O \leq$ 100 ma)		typ.	11	15	20	25	30	40	mV
Quiescent current	I_Q	typ.	3	3	3	3.1	3.1	3.1	mA
Output noise voltage (10 Hz to 10 kHz)	v_n	typ.	40	60	80	90	150	200	µV
Ripple rejection (120 Hz)	$\frac{\Delta V_{IN}}{\Delta V_{OUT}}$	typ.	60	55	52	49	46	43	dB
Dropout voltage	$V_{IN}-V_{OUT}$	typ.	1.7	1.7	1.7	1.7	1.7	1.7	V

When used as a zener diode/resistor combination replacement, the LM78 LXX usually results in an effective output impedance improvement of two orders of magnitude, and lower quiescent current. The regulators are available in the metal TO-5 (H) or in plastic TO-92 (Z) encapsulation. With adequate heat sinking the regulator can deliver 100 mA output current. Current limiting is included to limit the peak output current to a safe value. Safe area protection for the output transistor is provided to limit internal power dissipation. If internal power dissipation becomes too high for the heat sinking provided, the thermal shutdown circuit takes over preventing the IC from overheating.

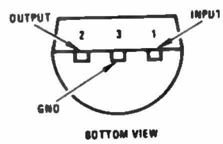
The series features output voltage tolerances of approximately 5% (LM-78LXXAC) and approximately 10% (LM78LXXC) over the temperature range.

LM3910/LM340L series

TO-39 (H)



TO-92 (Z)



LM3910, LM340L series (5, 6, 8, 10, 12, 15, 18, 24 V)

Output voltage	V_{OUT}		5	6	8	10	12	15	18	24 V
Input voltage	V_{IN}	typ.	10	11	14	16	19	23	27	33 V
	absolute max.		35	35	35	35	35	35	35	40 V
(for $V_{OUT} \pm 2\%$)		min.	7	8	10	12	14	17	20	27 V
		max.	20	21	23	25	27	30	33	38 V
Output current	I_{OUT}	typ.	40	40	40	40	40	40	40	40 mA
Junction temperature	T_J	max.	150	150	150	150	150	150	150	150 °C/W
Thermal resistance	T039	typ.	140	140	140	140	140	140	140	140 °C/W
	T092	typ.	180	180	180	180	180	180	180	180 °C/W
(junction to case)	T039	typ.	40	40	40	40	40	40	40	40 °C/W
Line regulation (I_O = 40 mA)	ΔV_O	typ.	18	20	20	25	30	37	45	60 mV
(I_O = 100 mA)		typ.	18	20	20	25	30	37	45	60 mW
Load regulation (I_O = 40 mA)		typ.	5	6	8	10	10	12	15	20 mV
(I_O = 100 mA)		typ.	11	13	15	20	20	25	30	40 mV
Quiescent current	I_Q	typ.	3	3	3	3	3	3.1	3.1	3.1 mA
Output noise voltage (10 Hz to 10 kHz)	v_n	typ.	40	50	60	70	80	90	150	200μV
Ripple rejection (120 Hz)	$\frac{\Delta V_{IN}}{\Delta V_{OUT}}$	typ.	62	60	58	57	54	52	50	48μV
Dropout voltage	V_{IN}-V_{OUT}	typ.	1.7	1.7	1.7	1.7	1.7	1.7	1.7	1.7 V

The LM3910 is an improved version o the LM78LXX series with a tighte output voltage tolerance, higher rippl rejection, better regulation and lowe quiescent current. The LM3910 regulat ors have approximately 2% Vou specification, 0.04%/V line regulation and 0.4% load regulation.

The LM3910 is available in the lov profile metal TO39 (H) packages anc the plastic TO-92 (Z) packages. Witl adequate heat sinking the regulato can deliver 100mA output current Current limiting is included to limit th peak output current to a safe value Safe area protection for the outpu transistor is provided to limit interna power dissipation. If internal powe dissipation becomes too high for th heat sinking provided, the therma shutdown circuit takes over preventin the IC from overheating.

Heat sink selection for a voltage regulator

Compute total thermal resistance

Determine the total thermal resistance, junction to ambient, necessary to maintain steady state T_J below the maximum value.

$$\theta_{JA(TOT)} = \frac{T_J - T_A}{P_D} \, ^{\circ}\mathrm{C/W}$$

Under the short circuit conditions, the internal thermal shutdown will limit T_J to about 175 ± 15°C. Although this protects the device, prolonged operation at such temperatures can adversely affect device reliability.

Determine if heat sink is required

Refer to the thermal resistance, θ_{JC} and θ_{JA}, columns of the data sheet summary.

a) $\theta_{JA(TOT)} > \theta_{JC}$ must be met, otherwise a higher wattage device must be used or a boost circuit employed.

b) If $\theta_{JA(TOT)} > \theta_{JA}$, a heat sink is not required

c) If $\theta_{JC} < \theta_{JA(TOT)} < \theta_{JA}$, a heat sink is required.

Select a heat sink

Choose a suitable heat sink from manufacturers' specification data. The necessary conditions are that θ_{JA} (TOT and $\theta'_{JA(TOT)}$ be less than θ_{JA}.

The total thermal resistance is tha from junction to case plus that fron case to ambient or sink to ambien (neglecting that from case to sink which is small).

$$\theta_{JA(TOT)} \approx \theta_{JC} + \theta_{SA} \, ^{\circ}\mathrm{C}$$

Check the input ripple and inpu variations

Insure that full-load $V_{IN(MIN)}$ does no allow V_{IN}-V_{OUT} to fall below th dropout voltage of about 2 V. Insur that no-load $V_{IN(MAX)}$ does not exceed the value listed on the data sheets.

When not to use a three-terminal voltage regulator

If you consider the voltage regulator as a black box you would not normally have any trouble, but there are occasions when these devices take off — here's the reason why.

The voltage regulator is essentially an amplifier with negative feedback. In this situation the gain varies with the load on the output. If the load is reactive then the feedback can become positive at particular frequencies — and the regulator breaks into oscillation. These devices should be treated with the same care one would give to an audio amplifier — they work with the loads specified but you should think twice about using anything unusual (like long wires in burglar alarms).

If you feel that it is necessary to adjust the phase of the feedback signal you should not be using a three-terminal regulator; there are many other regulators available which enable you to pu a calculated value of capacitance in th feedback path.

Manufacturers recommend puting 0.1 μF capacitor across the output o their regulators and this is usually sufficient precaution against oscillation You should always have two connections to the common terminal of th regulator — if you try to use one wire t carry both input and output current you are asking for trouble.

Intersil ICL7106/7107 digital panel meter IC

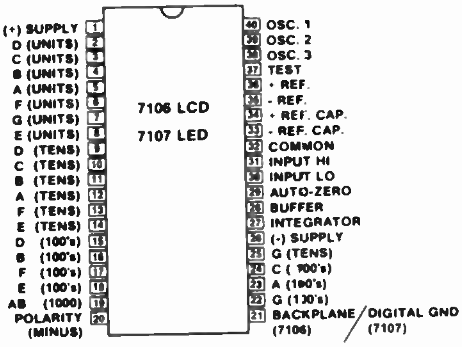


Fig. 1. Pin Configuration.

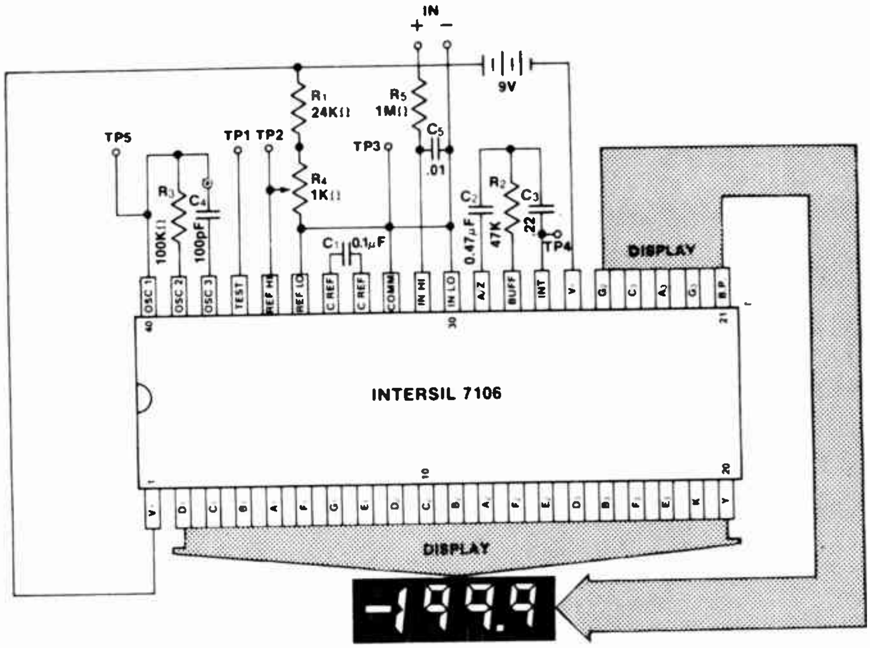


Fig. 3. LCD Digital Panel Meter Using ICL7106.

THE ICL7106 and 7107 are high performance, low power, CMOS 3½ digit A/D converters that contain all the necessary active devices on a single monolithic IC. Each has parallel seven-segment outputs which are ideal for use in a digital panel meter. The ICL7106 will directly drive a liquid crystal display including the backplane drive. The ICL7107 will directly drive instrument size LEDs without buffering. With seven passive components, display and power supply, the system forms a complete digital voltmeter with automatic zero connection and polarity (see figs. 3 and 4).

Both ICs use the time-proven dual slope integration technique with all its advantages, i.e. non-critical components, high noise rejection, non-critical clock frequency and almost perfect differential linearity. Both the ICL7106 and 7107 can be used not only with its internal reference, but true ratiometric reading applications may also be accomplished over a full scale input range of 199.9 mV to 1.999 V.

The accuracy of conversion is guaranteed to plus or minus 1 count over the entire plus or minus 2000 counts and the auto-zero facility provides a guaranteed zero reading for 0 volts input. However, the chip does provide a true polarity output at low voltages for null detection. Both chips have an on-board clock and reference circuitry, as well as overrange detection.

Displays and DPs

The additional components required to build a DPM are a display (either LCD or LED), 4 resistors, 4 capacitors, and an input filter if required. Liquid crystal displays become polarised and damaged if a DC voltage is continuously applied to them, so they must be driven with an AC signal. To turn on a segment, a waveform 180 degrees out of phase with the backplane drive (but of equal amplitude) is applied to that segment. The 7106 generates the segment drive waveform for all digits internally, but does not generate segment drive for the ▶

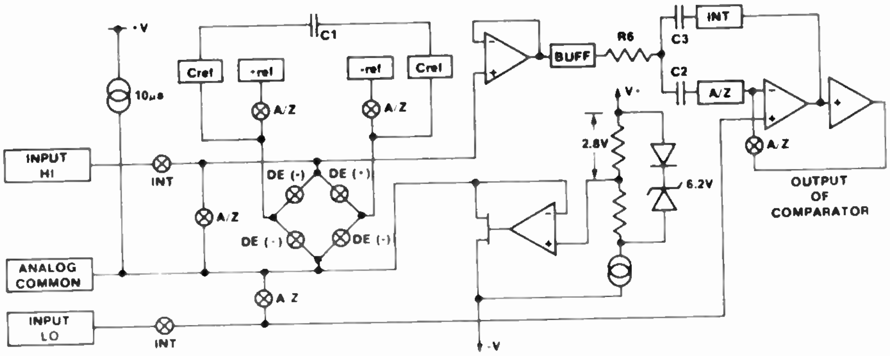


Fig. 2. Analogue Section Block Diagram.

ICL7106/7107

decimal point. This must be done using an inverter or exclusive-OR logic (see figs. 5 and 6). For use with LED displays the 7107 pull-down FETs will sink about 8 mA per segment, which produces a bright display suitable for almost any indoor application. A fixed decimal point can be turned on by tying the appropriate cathode to ground through a 150 ohm resistor.

Capacitors

The integration capacitor should be a low dielectric-loss type, such as a polypropylene. Mylar capacitors are suitable for the reference and auto-zero capacitors.

The Clock

The chip carries the active parts of an RC oscillator which runs at about 48 kHz and is divided by 4 for use as the system clock. The integration period (1000 clock pulses) is therefore 83.3 ms. Each conversion requires 4000 clock pulses, i.e. 3 readings per second. For optimum 50 Hz line frequency rejection, the clock should be set to a multiple of 50 Hz, e.g. 50 kHz.

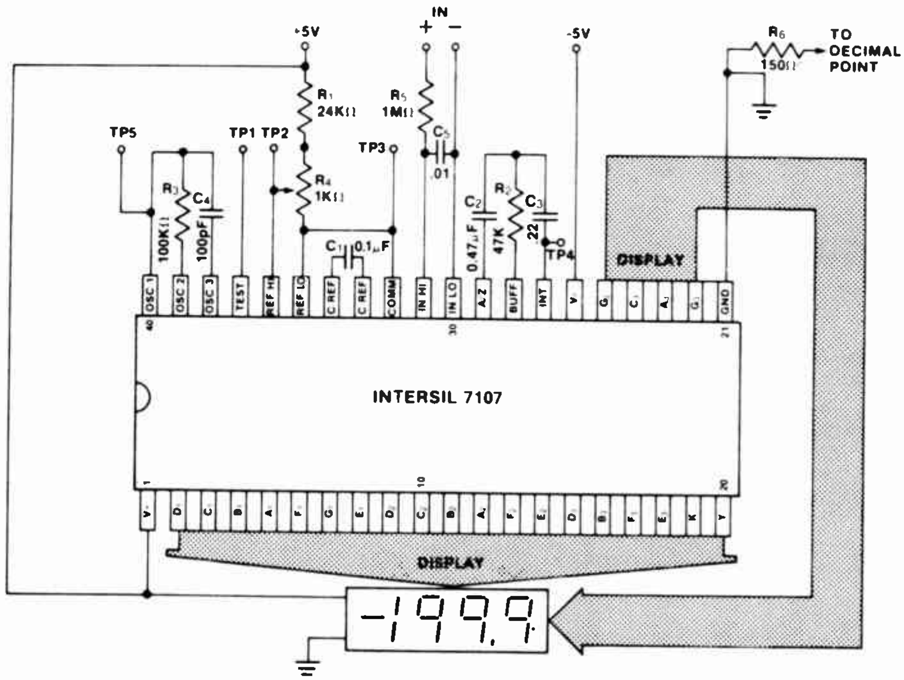


Fig. 4. LED Digital Panel Meter Using ICL7107.

The Reference

For 200.0 mV full scale, the voltage applied between REF Hi and REF Lo should be set at 100.0 mV. For 2.000 V full scale, this should be 1.000 V. The reference inputs are floating, and the only restriction on the applied voltage is that it should lie in the range V– to V+.

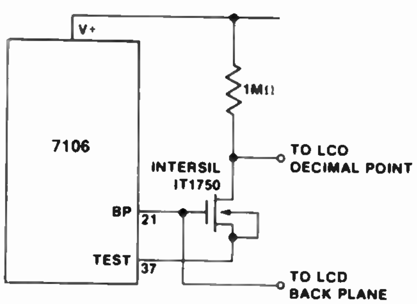


Fig. 5. Simple Inverter for fixed decimal point

For many applications, the internal reference of 2.8 V between V+ and COMMON is adequate, but power dissipation in the 7107 LED version can wreck this. However, an external reference can be added as shown in fig. 7.

Power Supplies

The 7106 will run from a single 5 to 12 V supply. If INPUT Lo is shorted to COMMON, this will cause V+ to sit 2.8 V positive with respect to INPUT Lo, and V– at 6.2 V negative with respect to INPUT Lo.

The 7107 requires dual supplies, +4.5 to +6 V and –3 to –6 V at 1 mA. A negative supply may be derived from +5 V using the circuit given in fig.8.

Further Information

Evaluation kits for the 7106 and 7107 are supplied with a data sheet and application note. In addition, Intersil produce three other Application Bulletins: A016 'Selecting A/D Converters', A017, 'The Integrating A/D Converter', and A018 'Do's and Dont's of Applying A/D Converters'.

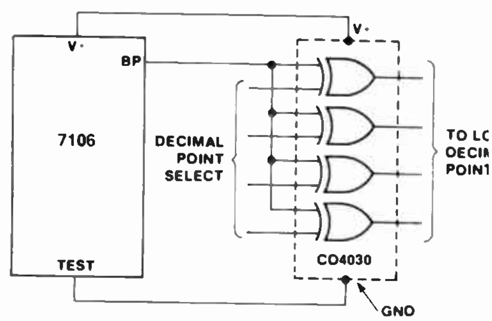
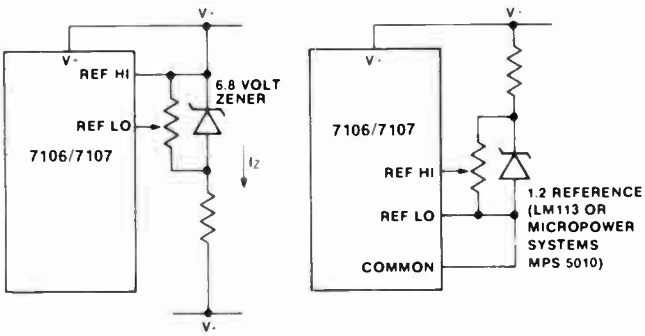


Fig. 6. Exclusive-OR gate for DP drive.



(a) (b)

Fig. 7. Using an external reference.

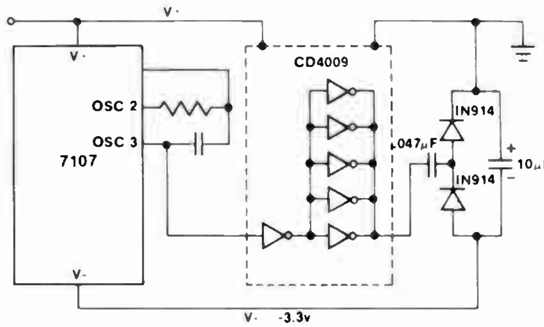


Fig. 8. Generating a negative supply from +5 V.

Electrical Specifications: ICL7106/7107

Full Scale Voltage Rating	∓200 mV (5.0 V min V+ to V–) ∓2.0 V (6.0 V min V+ to V–)
Full Scale Digital Range	∓2000 counts
Accuracy with external reference 10° to 50° C	<½ count
Noise referred to input	15 μV typical
Input circuit	Differential
Input Bias Current	2 pA
Input Impedance	>1 TΩ
Reference (Internal)	2.8 V, referenced to V+ Temperature Coefficient 100 ppm typical
Conversion Characteristic	Dual Slope with Auto-zero Integrating Time = 1000 counts Reference Time = 0 – 2000 counts Auto-zero time = 1000 + 2000 – Ref. Time

Recommended External Components:

	200 mV full scale	2 V full scale
Integrating Cap (C3)	0.22 μF	0.22 μF
AZ Cap (C2)	0.47 μF	0.047 μF
Ref. Cap (C1)	0.1 μF	0.1 μF
Clock Cap (C4)	100 pF	100 pF
Integrator Resistor (R6)	47 kΩ	470 kΩ
Clock Resistor (R3)	100 kΩ	100 kΩ

Clock Frequency	48 kHz, internally divided by 4
Power Requirements	LCD: 1 mA at 4.5 – 6 V LED: 1 mA at 4.5 – 6 V, plus LED current
Read Rate	Accurate from .1 to 15 readings per second

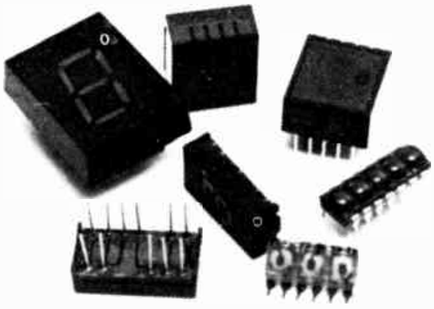
LED displays

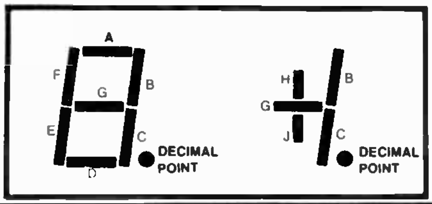
ABOUT THE ONLY feature common to the ranges of displays described in this Data Sheet is the way in which the various manufacturers identify the segments.

The standard method for doing this is shown below. We have deliberately excluded the 'overflow' type of L.E.D. display, in order to provide a better selection of normal types in the space available to us.

Calculated Omission

There is another type of L.E.D. display now becoming more popular in general usage. This is the calculator display, of a type personified by the HP device shown here. We hope to deal with these more fully at a later date. Generally these types use very low power, being, readable at about 100μA and with a varying number of digits, usually eight or ten.





The FND 507 is a common anod version of the FND 500, and as such car be used to replace a TIL 321.

Inclusion

Now that we've told you what isn't in here, perhaps we should explain what we have covered. Each display is described in a standard manner, using the same form of presentation for the relevant technical data. This is to facilitate easy comparison and subsequent selection.

Prices vary enormously from supplier to supplier, so we have not tried to give a definite price, just an indication Don't be mis-LED, some market segments might well display lower prices!

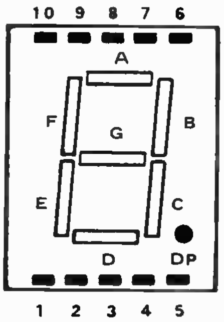
FND 500/507 — FAIRCHILD

The FND 500 is a common cathode display with an integral red filter. The decimal point is on the right-hand side of the device, which measures 15.3 mm by 16.5 mm high. This device is a pin for pin replacement to the Texas Instrument TIL322 display.

NOTE: While the FND500 is no longer manufactured, they are still in plentiful supply. In addition, Hewlett-Packard make a number of pin-compatible devices — the HDSP-5303 and HDSP5503 (high brightness). The Stanley NKR163 can be substituted, too. Both makes are physically narrower.

ELECTRICAL CHARACTERISTICS

DIGIT SIZE	0.5 ins
COLOUR	red
AVERAGE FWD CURRENT/SEGMENT	25mA
FORWARD VOLTAGE	1.7V
MIN. REV. BREAKDOWN VOLTAGE	3.0V
MAX. REV. CURRENT	100uA
LIGHT INTENSITY PER SEGMENT	600ucd
MAX. POWER DISSIPATION	400mW



PIN OUT – FND 500/507

1 Segment E	6 Segment B
2 Segment D	7 Segment A
3 Common	8 Common
4 Segment C	9 Segment F
5 Dec. point	10 Segment G

FND 500 . . . . common cathode
FND 507 . . . . common anode

DL 704/707

MONSANTO

A very common and widely available display, the 707 is the common anode version, with the 707R having a right-hand decimal point, as opposed to the standard left decimal on the 704 and 707. The 704 is thus a common cathode device.

ELECTRICAL CHARACTERISTICS

DIGIT SIZE	0.3 ins
COLOUR	YELLOW/RED/ORANGE
AVERAGE FWD CURRENT/SEGMENT	25mA
FORWARD VOLTAGE	2.5/1.6/1.6V
MIN. REV. BREAKDOWN VOLTAGE	3.0V
MAX. REV. CURRENT	100uA
LIGHT INTENSITY PER SEGMENT	320ucd
MAX. POWER DISSIPATION	500mW

PIN OUT–DL707/707R

1 Segment A	8 Segment D
2 Segment F	9 Anode
3 Anode	10 Segment C
4 NC	11 Segment G
5 NC	12 NC
6 Dec. point	13 Segment B
7 Segment E	14 Anode

PIN OUT–DL 704

1 Segment F	8 Segment C
2 Segment G	9 Dec. point
3 NC	10 NC
4 Cathode	11 NC
5 NC	12 Cathode
6 Segment E	13 Segment B
7 Segment D	14 Segment A

DL 747/750

MONSANTO

A 'Jumbo version' of the 707 and 704 devices. Widely available. Identify the common anode 747 by the missing pins – 1,9, 10 and 18.

The 750 is in full possession of its pins, and is common cathode. Decimal point is right-handed.

ELECTRICAL CHARACTERISTICS

DIGIT SIZE	0.6 ins
COLOUR	RED
AVERAGE FWD CURRENT/SEGMENT	25mA
FORWARD VOLTAGE	2.4V
MIN. REV. BREAKDOWN VOLTAGE	6.0V
MAX. REV. CURRENT	100uA
LIGHT INTENSITY PER SEGMENT	600ucd
MAX. POWER DISSIPATION	960mW

PIN OUTS–DL747/750

1 NC	10 NC
2 Segment A	11 Segment D
3 Segment F	12 Common
4 Common	13 Segment C
5 Segment E	14 Segment G
6 Common	15 Segment B
7 Dec. point	16 NC
8 NC	17 Common
9 NC	18 NC

DL747 . . . . common anode
DL750 . . . . common cathode
Pins 1, 9, 10, 18, omitted from 747

TIL RANGE — TEXAS

A uniform range of large displays, with red, green or amber encapsulation. No filters are needed, and a wide viewing angle is possible. Within defined categories, the devices are matched for luminous intensity. These can also act as direct replacements for the Fairchild FND500/507 duet.

PIN OUTS

TIL321/323/325

As FND 507. Direct replacement.

PIN OUTS—TIL322/324/326
As FND 500. Direct replacement.

ELECTRICAL CHARACTERISTICS

	321/322
DIGIT SIZE	015ins
COLOUR	RED
AVERAGE FWD CURRENT/SEGMENT	20mA
FORWARD VOLTAGE	1.7V
MIN. REV. BREAKDOWN VOLTAGE	3V
MAX. REV. CURRENT	100uA
LIGHT INTENSITY PER SEGMENT	600ucd
MAX. POWER DISSIPATION	300mW

ELECTRICAL CHARACTERISTICS

	323/324
DIGIT SIZE	0.5ins
COLOUR	GREEN
AVERAGE FWD CURRENT/SEGMENT	20mA
FORWARD VOLTAGE	2.5V
MIN. REV. BREAKDOWN VOLTAGE	3V
MAX. REV. CURRENT	100uA
LIGHT INTENSITY PER SEGMENT	320ucd
MAX. POWER DISSIPATION	600mW

ELECTRICAL CHARACTERISTICS

	325/26
DIGIT SIZE	0.5ins
COLOUR	AMBER
AVERAGE FWD CURRENT/SEGMENT	20mA
FORWARD VOLTAGE	2.5V
MIN. REV. BREAKDOWN VOLTAGE	3V
MAX. REV. CURRENT	100uA
LIGHT INTENSITY PER SEGMENT	340ucd
MAX. POWER DISSIPATION	400mW

XAN 352/4 — XCITON

These two come from what is the largest range of displays available. Xciton make big play of having all devices brighter than the competition, and a list of equivalents from their range for most of the others. These two are common cathode (XAN 354) and common anode (352) 0.3" numerics, using high efficiency GaAsP.

PIN OUT—XAN 352

1 Segment A	8 Segment D
2 Segment F	9 NC
3 Anode	10 Segment C
4 Omitted	11 Segment G
5 Omitted	12 Omitted
6 Dec. point	13 Segment B
7 Segment E	14 Anode

Agent: R. & D. Electronics Pty. Ltd., 23 Burwood Road, Burwood, 3125 Vic. Ph. 288 8262.

PIN OUT—XAN 354

1 Segment F	8 Segment C
2 Segment G	9 Dec. point
3 Omitted	10 Omitted
4 Cathode	11 Omitted
5 Omitted	12 Cathode
6 Segment E	13 Segment B
7 Segment D	14 Segment A

ELECTRICAL CHARACTERISTICS

DIGIT SIZE	0.3ins
COLOUR	GREEN
AVERAGE FWD CURRENT/SEGMENT	25mA
FORWARD VOLTAGE	2.0V
MIN. REV. BREAKDOWN VOLTAGE	5V
MAX. REV. CURRENT	100uA
LIGHT INTENSITY PER SEGMENT	450ucd
MAX. POWER DISSAPATION	400mW

XAN 650 SERIES — XCITON

ELECTRICAL CHARACTERISTICS 82/84

DIGIT SIZE	0.6ins
COLOUR	YELLOW
AVERAGE FWD CURRENT/SEGMENT	25mA
FORWARD VOLTAGE	2.2V
MIN. REV. BREAKDOWN VOLTAGE	3.0V
MAX.REV. CURRENT	100uA
LIGHT INTENSITY PER SEGMENT	700ucd
MAX. POWER DISSIPATION	400mW

PIN OUT XAN 652/654/682/684

1 Segment A	8 Segment D
2 Segment F	9 Common
3 Common	10 Segment C
4 Segment E	11 Segment G
5 Omitted	12 Segment B
6 Dec. point	13 Omitted
7 Omitted	14 Common

XAN 684/654. . . . . . common cathode
XAN 682/652. . . . . . . common anode

ELECTRICAL CHARACTERISTICS 52/54

DIGIT SIZE	0.6ins
COLOUR	GREEN
AVERAGE FWD CURRENT/SEGMENT	25mA
FORWARD VOLTAGE	2.0V
MIN. REV. BREAKDOWN VOLTAGE	3V
MAX. REV. CURRENT	100uA
LIGHT INTENSITY PER SEGMENT	2000ucd
MAX. POWER DISSIPATION	350mW

7400 SERIES — HEWLETT-PACKARD

The 7400 series are 2.79mm GaSP numeric indicators, packaged in end stackable DIL casings. They are readable at 500μA per segment, and constructed for strobed operation in such a way that less lead connections are needed. A lens magnifier is fitted, with a good viewing angle.

ELECTRICAL CHARACTERISTICS

DIGIT SIZE	(magnifier) 0.11ins
COLOUR	RED
AVERAGE FWD CURRENT/SEGMENT	5mA
FORWARD VOLTAGE	1.6V
MIN. REV. BREAKDOWN VOLTAGE	5V
MAX. REV. CURRENT	100uA
LIGHT INTENSITY PER SEGMENT	20ucd
MAX. POWER DISSIPATION	80mW

PIN OUT HP7402/7412

1 NC	8 Segment D
2 Segment C	9 Segment F
3 Segment C	10 Cathode
4 Cathode	11 Segment B
5 Dec. point	12 Segment A
6 Cathode	13 Omitted
7 Segment G	14 Omitted

PIN OUT HP7404/7414

1 Cathode	8 Segment D
2 Segment E	9 Segment F
3 Segment C	10 Cathode
4 Cathode	11 Segment B
5 Dec. point	12 Segment A
6 Cathode	13 Omitted
7 Segment G	14 Omitted

PIN OUT HP 7403/7413

1 Cathode	8 Segment D
2 Segment E	9 Segment F
3 Segment C	10 Cathode
4 Cathode	11 Segment B
5 Dec. point	12 Segment A
6 NC	13 Omitted
7 Segment G	14 Omitted

PIN OUT HP7405/7415

1 Cathode	8 Segment 9
2 Segment E	9 Cathode
3 Segment C	10 Segment F
4 Cathode	11 NC
5 Dec. point	12 Segment B
6 Segment D	13 Cathode
7 Cathode	14 Segment A

Digits per Cluster	Center Decimal Point	Right Decimal Point
3 (right)	5082-7402	5082-7412
3 (left)	5082-7403	5082-7413
4	5082-7404	5082-7414
5	5082-7405	5082-7414

Intersil ICM7216A/B/C/D

ICM7216A Universal Counter, Drives Common Anode LED's
ICM7216B Universal Counter, Drives Common Cathode LED's
ICM7216C Frequency Counter, Drives Common Anode LED's
ICM7216D Frequency Counter, Drives Common Cathode LED's

FEATURES

ICM7216A AND B

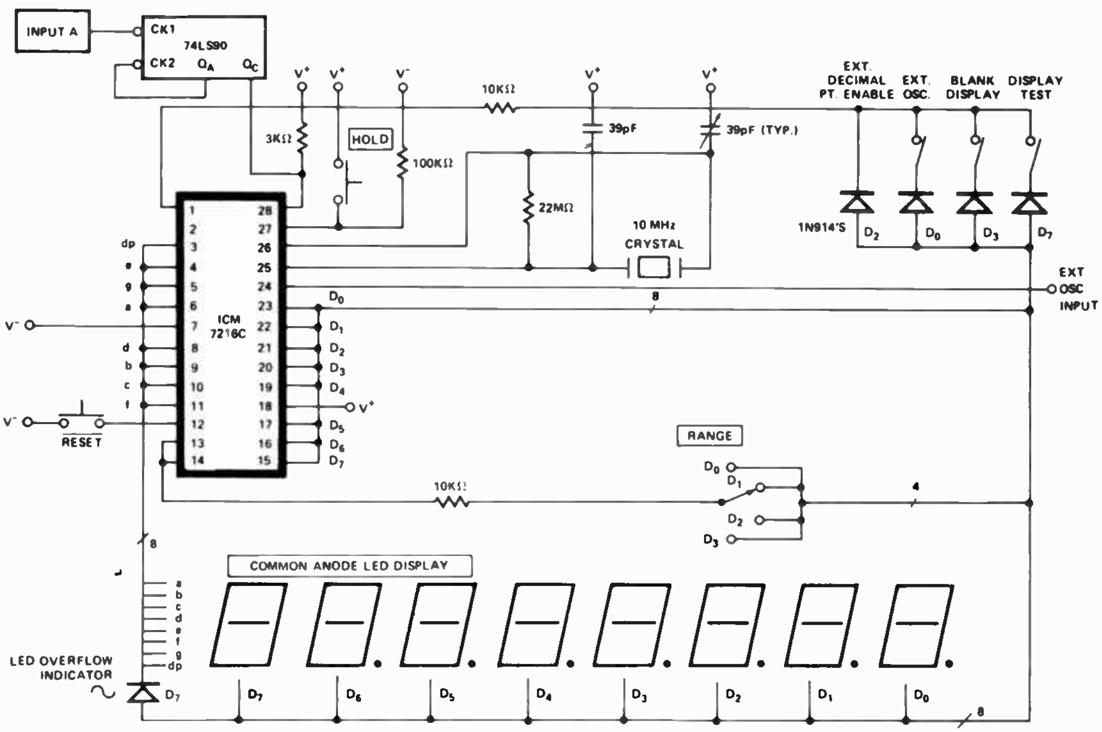
- Functions as a Frequency Counter, Period Counter, Unit Counter, Frequency Ratio Counter or Time Interval Counter
- Four Internal Gate Times:
  0.01 sec, 0.1 sec, 1 sec, 10 sec in Frequency Counter Mode
- 1 Cycle, 10 Cycles, 100 Cycles, 1000 Cycles in Period, Frequency Ratio and Time Interval Modes
- Measures Frequencies from DC to 10 MHz
- Measures Period from 0.5μ sec to 10 sec

ICM7216C AND D

- Functions as a Frequency Counter. Measures Frequencies from DC to 10 MHz
- Decimal Point and Leading Zero Blanking May be Externally Selected

ALL VERSIONS:

- Eight Digit Multiplexed LED Outputs
- Output Drivers will Directly Drive Both Digits and Segments of Large LED Displays. Both Common Anode and Common Cathode Versions are Available
- Single Nominal 5V Supply Required
- Stable High Frequency Oscillator, Uses Either 1 MHz or 10 MHz Crystal
- Internally Generated Multiplex Timing with Interdigit Blanking, Leading Zero Blanking and Overflow indication
- Decimal Point and Leading Zero Blanking Controlled Directly by the Chip
- Display Off Mode Turns Off Display and Puts Chip into Low Power Mode
- Hold and Reset Inputs for Additional Flexibility
- Test Speedup Function Included
- All Terminals Protected Against Static Discharge



100MHz Frequency Counter

GENERAL DESCRIPTION

The ICM7216A and B are fully integrated Universal Counters and LED display drivers. They combine a high frequency oscillator, a decade timebase counter, an 8 decade data counter and latches, a 7 segment decoder, digit multiplexers and 8 segment and 8 digit drivers which can directly drive large LED displays. The counter inputs have a maximum frequency of 10 MHz in frequency and unit counter modes and 2 MHz in other modes. Both inputs are digital inputs. In many applications, amplification and level shifting will be required to obtain proper digital signals for these inputs.

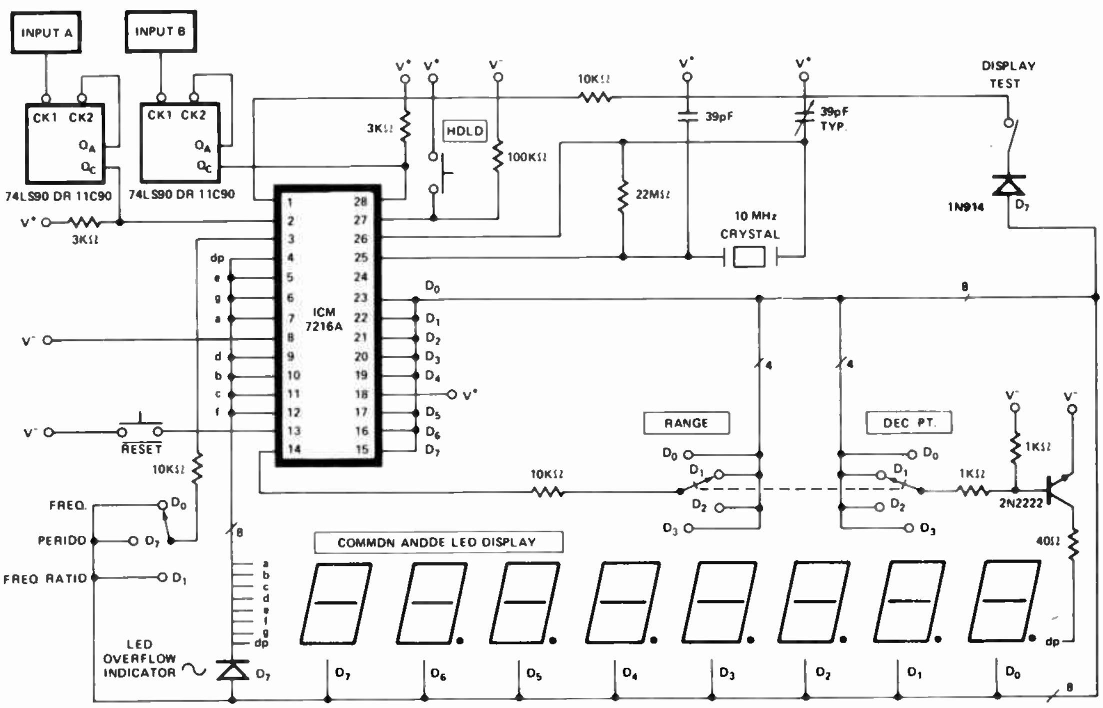
The ICM7216A and B can function as a frequency counter, period counter, frequency ratio (fA/fB) counter, time interval counter or as a totalizing counter. The counter uses either a 10 MHz or 1 MHz quartz crystal timebase. An external timebase input is also provided. For period and time interval, the 10 MHz timebase gives a 0.1 μsec resolution. In period average and time interval average, the resolution can be in the nanosecond range. In the frequency mode, the user can select accumulation times of 0.01 sec, 0.1 sec, 1 sec and 10 sec. With a 10 sec accumulation time, the frequency can be displayed to an accuracy of 0.1 Hz in the least significant digit. There is 0.2 seconds between measurements in all ranges.

The ICM7216C and D function as frequency counters only, as described above.

All versions of the ICM7216 incorporate leading zero blanking. Frequency is displayed in KHz. In the ICM7216A and B, time is displayed in μsec. The display is multiplexed at 500Hz with a 12.5% duty dycle for each digit. The ICM7216A and C are designed for common anode display with typical peak segment currents of 25mA. The ICM7216B and D are designed for common cathode displays with typical peak segment currents of 12mA. In the display off mode, both digit drivers and segment drivers are turned off enabling the display to be used for other functions.

ABSOLUTE MAXIMUM RATINGS

Maximum Supply Voltage ($V^+ - V^-$) . . . . 6.5 Volts
Maximum Digit Output Current 400mA
Maximum Segment Output Current 60mA
Voltage On Any Input or
Output Terminal [1] . . . . V^+ + .3V to V^- – .3V
Maximum Power Dissipation at
70°C . . . . 1.0 Watts (ICM7216A & C)
0.5 Watts (ICM7216B & D)



100MHz Multifunction Counter

BFXX-35 power fets

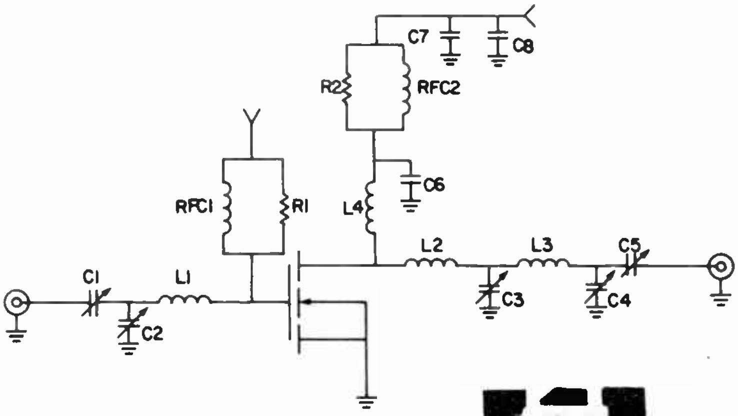
HIGH POWER VFETs will gradually supercede bipolar transistor technology in RF power amplifier applications in the next few years. Cost, on a dollar-per-watt basis, is rapidly decreasing on VFET devices, and when the overall cost of a complete amplifier – or amplifier chain – is considered, the circuit simplicity using VFETs brings cost benefits of its own.

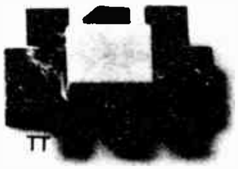
VFETs have a negative temperature coefficient and are not subject to thermal runaway. Severe load mismatch presents few problems to these devices in RF power amplifiers.

This range of VFET devices is from the Communications Transistor Corporation (CTC), represented in Australia by Ampec Engineering, 1 Wellington St., Rozelle NSW 2039 (02) 818-1166. The three devices presented here are N-channel types, suitable for HF and VHF amplifier applications (linear – class B, or saturated – class C), running from a 28 to 35 volt supply rail.

CAUTION

Users are cautioned that these devices use Beryllium Oxide ceramics in their construction. Any mechanical or chemical treatment of these ceramics which produces dust or fumes, even minute amounts, can be dangerous. Care should be taken to ensure that those who handle, use or dispose of these devices are aware of this. The package should never be crushed, filed, chipped, sawed, sanded, ground or come in contact with acid. In the event of failure, return to the manufacturer.



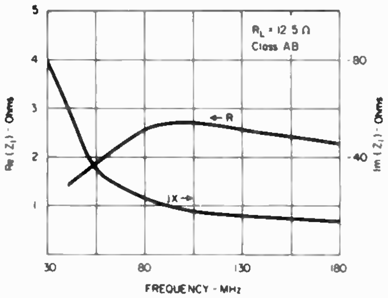


Package, 8F100-35

TEST CIRCUIT 175 MHz 100 Wout BF100-35

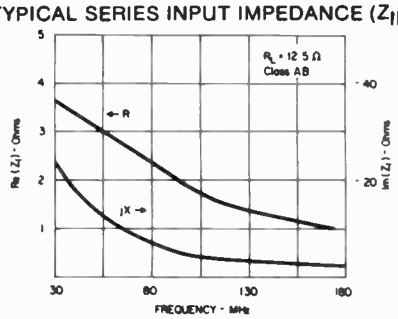
C1, C2, C4	5-70 p compressed mica
C3, C5	25-240 p compressed mica
C6	470 p ceramic chip
C7	100 n ceramic
C8	10μ electrolytic
L1	half turn No. 18 wire, 10 mm ID
L2	1 mm thick copper strap 5 mm wide, 16 mm long, 10 mm ID loop
L3	1½ turns No. 18 wire, 10 mm ID
L4	6 turns No. 18 wire, 10 mm ID
RFC1	6 turns No. 18 wire on Indian General F627-8 Q1 toroid.
RFC2	18 turns No. 18 wire on Micrometals T106-2, toroid with 15 R, 2 W resistor in parallel.

TYPICAL SERIES INPUT IMPEDANCE (Z_{IN})



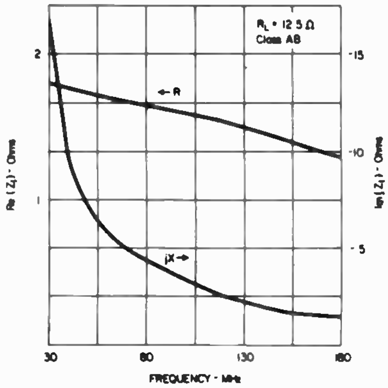
BF25-35

TYPICAL SERIES INPUT IMPEDANCE (Z_{IN})



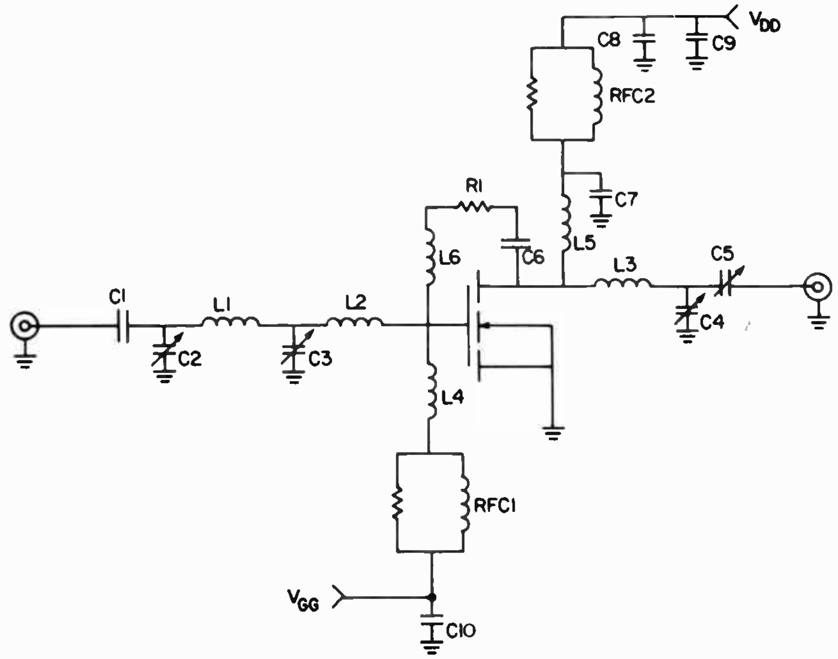
BF50-35

TYPICAL SERIES INPUT IMPEDANCE (Z_{IN})



BF100-35

Package, BF25-35 and BF50-35



TEST CIRCUIT 80 MHz 50Wout/BF50-35 , 25 Wout/BF25-35

C1	1 n ceramic chip	L5	6 turns No. 18 wire, 10 mm ID
C2	5-70 p compressed mica	L6	150 nH (0.15uH)
C3-C5	10-180 p compressed mica	RFC1	16 turns No. 22 wire on Indiana General T106-2 toroid with 15 R, ½ W resistor in parallel
C6	100 n ceramic		
C7	470 p ceramic chip		
C8, C10	10 n ceramic	RFC2	18 turns No. 18 wire on Indiana General T106-2 toroid with 15 R, 2 W resistor in parallel
C9	10u electrolytic		
L1	3 turns No. 18 wire, 10 mm ID		
L2, L3	2½ turns No. 18 wire, 10 mm ID	R1	150 R, ½ W
L4	330 nH (0.33 uH)		

	Symbol	BF25-35	BF50-35	BF100-35	units
Maximum power dissipation with 25°C case temperature		75	125	250	W
Maximum drain to source voltage	BV_{DSS}		65		V
Maximum gate to source voltage	BV_{GS}		25		V
Maximum drain current	I_D	5	10	16	A
Power output with f=175 MHz, V_{DS}=35 V	P_{out}	25	50	100	W
Power input under above conditions	P_{IN}	2.5	6	16	W
Drain efficiency under above conditions	η		50		%
ON-state drain current	$I_{D(ON)}$	3.5	5	10	A
Small signal forward transconductance	g_M	0.7	1.5	2.2	MHO
Drain-source capacitance with f=1 MHz, V_{DS}=28V and V_{GS}=0 V	C_{DS}	37	77	170	pF

555 & 556 timing circuits

QUICK REFERENCE DATA
555 TIMER

Absolute maximum ratings		
Supply voltage	max.	18 V
Power dissipation	max.	600 mW
Characteristics		
(25°C, V_{CC} 5 to 15 V)		
Supply voltage		4.5 to 15 V
Supply current (low state) (V_{CC} = 5 V)	typ.	3 mA
(V_{CC} = 15 V)	typ.	10 mA
Timing error, initial accuracy	typ.	1%
Timing error, temperature drift	typ.	50 ppm/°C
Timing error, drift with supply voltage	typ.	0.1 %/V
Threshold voltage	typ.	1/3 V_{CC}
Control voltage level (V_{CC} = 15 V)	typ.	10 V
(V_{CC} = 5 V)	typ.	3.3 V
Output voltage drop (LOW)		
V_{CC} = 15 V; I_{SINK} = 10 mA	typ.	0.1 V
I_{SINK} = 50 mA	typ.	0.4 V
I_{SINK} = 100 mA	typ.	2 V
V_{CC} = 5 V; I_{SINK} = 5 mA	typ.	0.25 V
Output voltage drop (HIGH)		
V_{CC} = 15 V; I_{SOURCE} = 200 mA	typ.	12.5 V
I_{SOURCE} = 100 mA	typ.	13.3 V
V_{CC} = 5 V	typ.	3.3 V
Risetime/falltime of output	typ.	100 ns

The 555 timer is available from many manufacturers under codings like NE555, µA555, LM555, MC1455 and MC1555. Recent ETIs have carried ads pricing the device at around 60c to 90c.

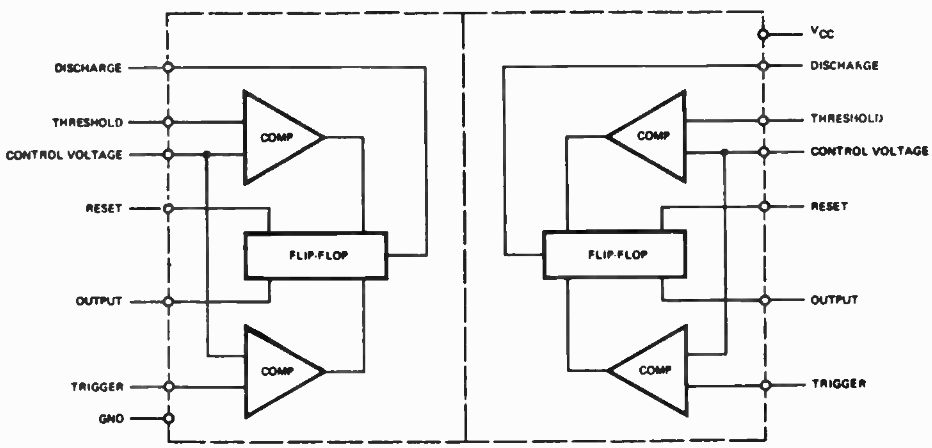
The IC is a stable controller which produces very accurate time delays or rectangular-waveform oscillations. One external capacitor and resistor set the time delay, and in the oscillator mode one further resistor is all that is needed to give control of frequency and duty cycle.

A trigger signal sets an internal flip-flop and starts the timing (the flip-flop immunises the device from further triggering). A reset signal can be applied to interrupt the timing cycle.

The output is capable of sinking or sourcing 200 mA to drive relays, indicators or further circuitry.

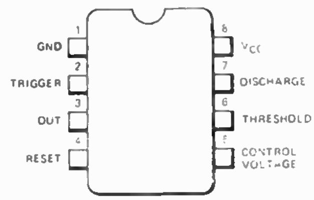
The 556 contains two 555s in one package, for sequential or multiple applications.

BLOCK DIAGRAM– 556

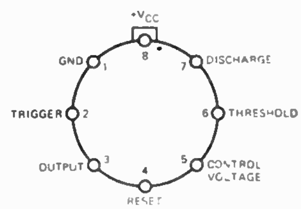


PINOUTS– 555

8-LEAD MINI DIP
(TOP VIEW)

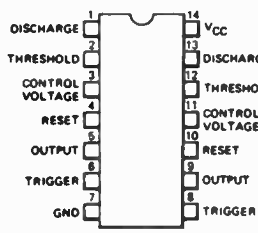


8-LEAD TO-100
(TOP VIEW)



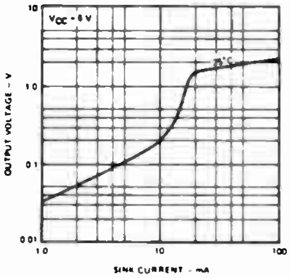
PINOUT– 556

14-LEAD DIP
(TOP VIEW)

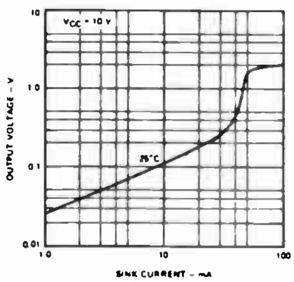


Notes

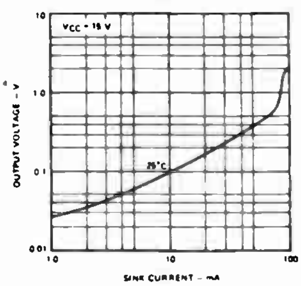
LOW OUTPUT VOLTAGE AS A FUNCTION OF OUTPUT SINK CURRENT



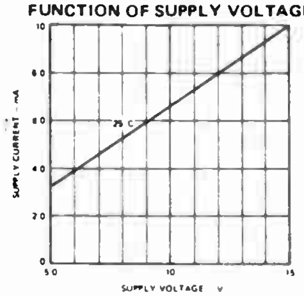
LOW OUTPUT VOLTAGE AS A FUNCTION OF OUTPUT SINK CURRENT



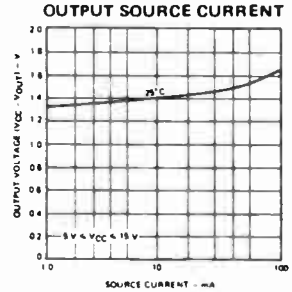
LOW OUTPUT VOLTAGE AS A FUNCTION OF OUTPUT SINK CURRENT



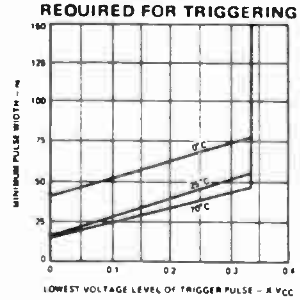
TOTAL SUPPLY CURRENT AS A FUNCTION OF SUPPLY VOLTAGE



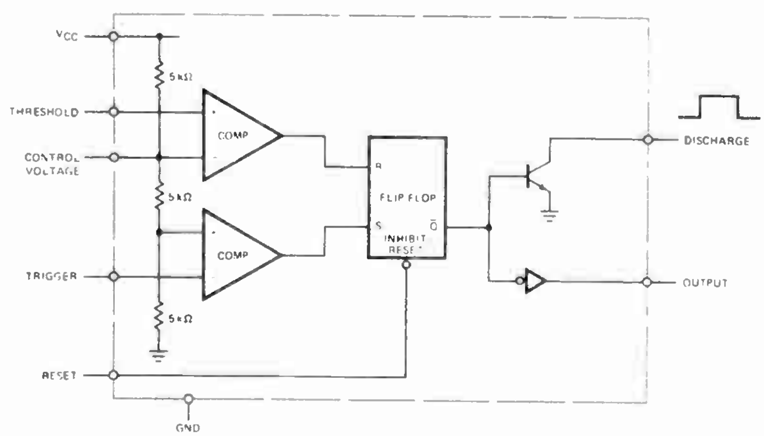
HIGH OUTPUT VOLTAGE AS A FUNCTION OF OUTPUT SOURCE CURRENT



MINIMUM PULSE WIDTH REQUIRED FOR TRIGGERING



BLOCK DIAGRAM— 555



555 & 556 timing circuits cont'

TYPICAL APPLICATIONS

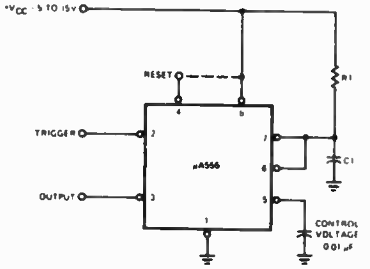
MONOSTABLE OPERATION

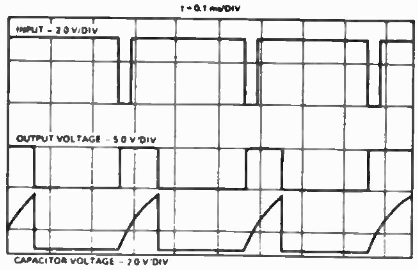
In the monostable mode, the timer functions as a one-shot. Referring to Figure 1 the external capacitor is initially held discharged by a transistor inside the timer.

When a negative trigger pulse is applied to lead 2, the flip-flop is set, releasing the short circuit across the external capacitor and drives the output HIGH. The voltage across the capacitor, increases exponentially with the time constant τ = R1C1. When the voltage across the capacitor equals 2/3 V_{CC}, the comparator resets the flip-flop which then discharges the capacitor rapidly and drives the output to its LOW state. Figure 2 shows the actual waveforms generated in this mode of operation.

The circuit triggers on a negative-going input signal when the level reaches 1/3 V_{CC}. Once triggered, the circuit remains in this state until the set time has elapsed, even if it is triggered again during this interval. The duration of the output HIGH state is given by t = 1.1 R1C1 and is easily determined by Figure 3. Notice that since the charge rate and the threshold level of the comparator are both directly proportional to supply voltage, the timing interval is independent of supply. Applying a negative pulse simultaneously to the Reset terminal (lead 4) and the Trigger terminal (lead 2) during the timing cycle discharges the external capacitor and causes the cycle to start over. The timing cycle now starts on the positive edge of the reset pulse. During the time the reset pulse is applied, the output is driven to its LOW state.

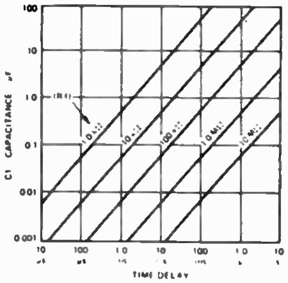
When Reset is not used, it should be tied high to avoid any possibility of false triggering.





R1 = 9.1 kΩ C1 = 0.01 μF RL = 1.0 kΩ

TIME DELAY AS A FUNCTION OF R1 AND C1



ASTABLE OPERATION

When the circuit is connected as shown in Figure 4 (leads 2 and 6 connected) it triggers itself and free runs as a multivibrator. The external capacitor charges through R1 and R2 and discharges through R2 only. Thus the duty cycle may be precisely set by the ratio of these two resistors.

In the astable mode of operation, C1 charges and discharges between 1/3 V_{CC} and 2/3 V_{CC}. As in the triggered mode, the charge and discharge times and therefore frequency are independent of the supply voltage.

Figure 5 shows actual waveforms generated in this mode of operation.

The charge time (output HIGH) is given by:

$$t_1 = 0.693\ (R1 + R2)\ C1$$

and the discharge time (output LOW) by:

$$t_2 = 0.693\ (R2)\ C1$$

Thus the total period T is given by:

$$T = t_1 + t_2 = 0.693\ (R1 + 2R2)\ C1$$

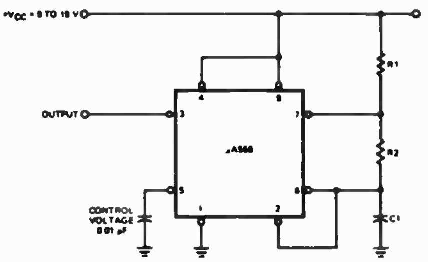
The frequency of oscillation is then:

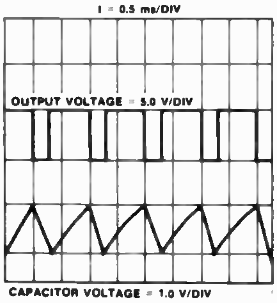
$$f = \frac{1}{T} = \frac{1.44}{(R1 + 2R2)\ C1}$$

and may be easily found by Figure 6.

The duty cycle is given by:

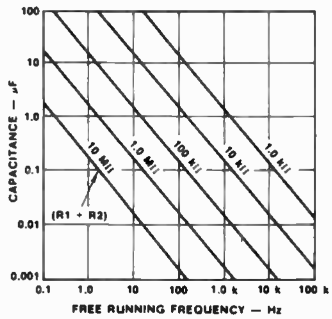
$$D = \frac{R2}{R1 + 2R2}$$





R1 = R2 = 4.8 kΩ, C1 = 0.1 μF, RL = 1 kΩ

Free Running Frequency vs R1, R2, and C1



APPLICATIONS– CIRCUITS

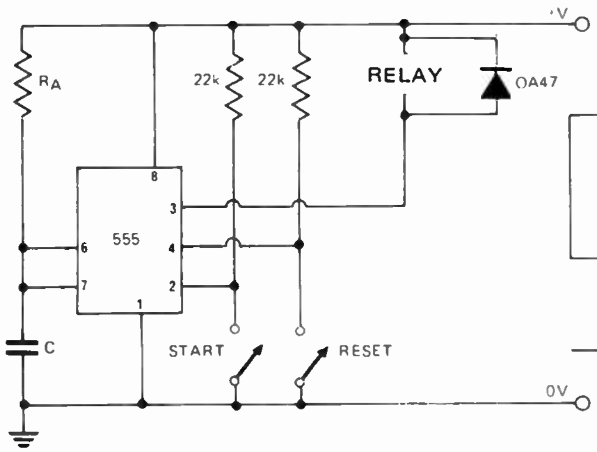


Fig. 4

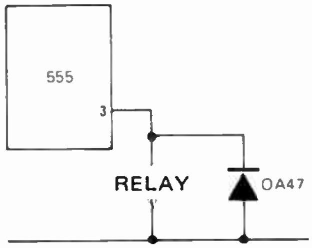


Fig. 5

Fig. 1. Using the 555 to drive a relay. The delay period should be greater than 0.1s and the relay should operate in the 555 supply voltage and draw not more than 200 mA. Here the relay is normally closed — it opens for the delay period. Since the current required by the trigger pin (2) is only 0.5 µA for 0.1 µs, it can be triggered by pick-up of the voltage transient produced by the relay coil when it switches off (the result being a relay which doesn't actually open at the end of the timing period). Figure 2 shows how to connect a diode across the relay to prevent this trouble, but note that the diode has to act pretty fast: gold-bonded germanium types (such as 0A47) are the best, silicon types (like the 1N914) are unsatisfactory.

Fig. 3. The simplest 555 astable circuit uses one resistor, one capacitor and the 555. The charging and discharging times are both approximately 0.7RC. A relay may be connected (with suitable diode) between pin 3 and either supply rail. Alternatively the output may be set at audio frequency to provide a square-wave for testing audio gear.

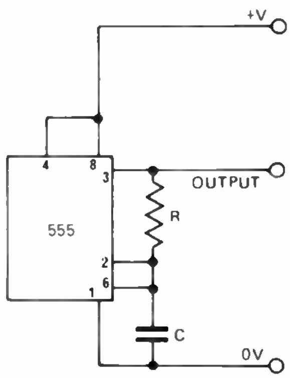


Fig. 9

Fig. 4. Using the 555 as a comparator, in this case as a photosensitive switch. The switch-on threshold is reached when the voltage at pin 2 falls to a third of the supply voltage. The second threshold is reached when the voltage rises to two-thirds of the supply voltage — the relay then opens. Adjusting the resistor R will change the light-level for a given cadmium sulphide photo cell. The photocell may be replaced by a thermistor or other device which changes in resistance.

Because the output switches in and off at different levels you get a useful hysteresis effect. This stops any unwanted switching on and off as the light level hovers about the threshold level.

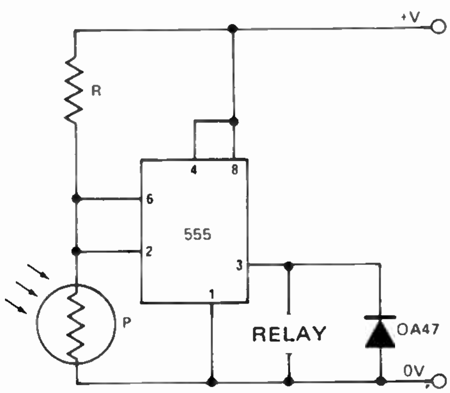


Fig. 10.

CMOS

This is a list of commonly-available CMOS in the 4000 and 74C00 series (plus one or two other similar types).

When you see a blob in front of a number this means that the device is available in LOCMOS. LOCMOS ICs are functionally identical to other CMOS but the outputs are buffered. This means that they cannot be operated in the linear mode (as amplifiers, etc). CMOS from Solid-State Scientific is also buffered.

Nor/nand

●4000	Dual 3-Input NOR Gate Plus Inverter
●4001	Quad 2 Input NOR Gate
●4002	Dual 4 Input NOR Gate
●4011	Quad 2 Input NAND Gate
●4012	Dual 4 Input NAND Gate
●4023	Triple 3 Input NAND Gate
●4025	Triple 3 Input NOR Gate
●4068	8-Input NAND Gate
●4078	8-Input NOR Gate
40107	Dual 2-Input NAND Buffer Driver
74C00	Quad 2-Input NAND Gate
74C02	Quad 2-Input NOR Gate
74C10	Triple 3-Input NAND Gate
74C20	Dual 4-Input NAND Gate
74C30	8-Input NAND Gate

Or/and

●4071	Quad 2 Input OR Gate
●4072	Dual 4 Input OR Gate
●4073	Triple 3-Input AND Gate
●4075	Triple 3 Input OR Gate
●4081	Quad 2 Input AND Gate
●4082	Dual 4-Input AND Gate
74C08	Quad 2-Input AND Gate
74C32	Quad 2-Input OR Gate

Buffers & inverters

●4007	Dual Complementary Pair Plus Inverter
4009	Hex Inverting Buffer
4010	Hex Buffer
●4041	Quad True Complement Buffer
●4049	Hex Inverting Buffer
●4050	Hex Buffer
●4069	Hex Inverter
●4502	Strobed Hex Inverter Buffer
4503	Hex Bus Driver
●40097	3-S Hex Non-Inverting Buffer
●40098	3-S Hex Inverting Buffer
40107	Dual 2-Input NAND Buffer Driver
74C04	Hex Inverter

Single level gates

74C901	Hex Inverting Buffer (TTL Interface)
74C902	Hex Buffer (TTL Interface)
74C903	Hex Inverting Buffer (MOS Interface)
74C904	Hex Buffer (MOS Interface)
74C906	Open Drain Buffer (Active Pull-down)
74C907	Open Drain Buffer (Active Pull-down)
74C908	Dual 30V Buffer
80C95	TRI-STATE Hex Buffer
80C96	TRI-STATE Hex Inverting Buffer
80C97	TRI-STATE Hex Buffer
80C98	TRI-STATE Hex Inverting Buffer
88C29	Quad Line Driver (Single Ended)
88C30	Dual Line Driver (Twisted Pa

Level shifter

40109	Quad Low-to-High Voltage Level Shifter

Multi-level gates

Multifunction & AOI

●4019	Quad AND/OR Select Gate
●4030	Quad EX OR Gate
4037	Triple AND/OR Bi-Phase Pairs
4048	Expandable 8-Input Gate
●4070	Quad Exclusive OR Gate
●4077	Quad Exclusive NOR Gate
●4085	Dual 2 Wide 2-Input AND/OR INVERT Gate
●4086	Expandable 4-wide 2-Input AOI
4507	Quad EX OR Gate (74C86)
74C86	Quad EX-OR Gate

Decoders & encoders

●4028	BCD to Decimal Decoder
●4514	4-Bit Latch/4 to 16 Line Decoder (High)
●4515	4-Bit Latch/4 to 16 Line Decoder (Low)
●4532	8-Input Priority Encoder
●4555	Dual 1-of-4 Decoder/Demultiplexer (Outputs High)
●4556	Dual 1-of-4 Decoder Demultiplexer (Outputs Low)
74C42	BCD-to-Decimal Decoder
74C48	BCD-to-7 Segment Decoder
74C154	4 to 16 Line Decoder
74C915	7-Segment to BCD Decoder
74C922	16 Key Keyboard Encoder
74C923	20 Key Keyboard Encoder

Schmidtt triggers

●4093	Quad 2-Input NAND Schmitt Trigger
40106	Hex Schmitt Trigger
4584	Hex Schmitt Trigger (74C14
74C14	Hex Schmitt Trigger
74C914	Hex Schmitt Trigger

Multivibrators

flip - flops & latches

●4013	Dual D Flip-Flop
●4027	Dual JK Flip-Flop
●4042	Quad D Latch
●4043	Quad 3 State NOR R/S Latch
●4044	Quad 3 State NAND R/S Lat
●4076	TRI-STATE Quad D Flip-Flc
4095	Gated J-K Non-Inverting Inputs
4096	Gated J-K Inverting and Non Inverting Inputs
4099	8-Bit Addressable Latch

Notes

●40174	Hex D Flip-Flop
●40175	Quad D Flip-Flop
4508	Dual 4-Bit Latch
●4724	8-Bit Address Latch
74C73	Dual JK Flip-Flop
74C74	Dual D Flip-Flop
74C76	Dual JK Flip-Flop
74C107	Dual JK Flip-Flop
74C173	TRI-STATE Quad D Flip-Flop
74C174	Hex D Flip-Flop
74C175	Quad D Flip-Flop

Astables & monostables

●4047	Monostable-Astable Multivibrator
4098	Dual Monostable Multivibrator
●4528	Dual Retriggerable/Resettable Monostable Multivibrator
74C221	Dual Monostable Multivibrator

Registers shift, storage & fifo

●4006	18-Bit S/R
●4014	8-Bit S/R
●4015	Dual 4-Bit S/R
●4021	8-Bit S/R
●4031	64-Bit S/R
4034	8-Bit S/R
●4035	4-Bit P-In P-Out S/R
4094	8-stage Shift-and-Store Bus Register
40100	32-Bit Left/Right S/R
40104	3-state 4-Bit Left/Right S/R
●40194	4-Bit Left/Right S/R
●40195	b-Bit Universal S/R
●4076	TRI-STATE Quad D Flip-Flop
4099	8-Bit Addressable Latch
40108	4x4 multiport register
40105	4-word 4-Bit FIFO Buffer
74C164	8-Bit S-In P-Out S/R
74C165	8-Bit P-In S-Out S/R
74C195	4-Bit Parallel S/R
74C95	4-Bit R-S L-S Register

Counters

●4017	Decade Counter Divider
●40108	Presettable Divide by "N" Counter
●4020	14-Bit Ripple Carry Binary Counter/Divider
●4022	Divide by 8 Counter Divider
●4024	7-Bit Binary Counter
●4029	Presettable Up/Down Counter
4040	12-Bit Binary/Ripple Counter
4045	21-stage Clock Timer
4059	Programmable Divide-by N Counter
4060	14-Stage Ripple Carry Binary
●4510	BCD Up/Down Counter
●4516	Binary Up/Down Counter
●4518	Dual BCD Up Counter
●4520	Dual Binary Up Counter
40102	BCD Presettable 8-Bit down Counter
40103	Binary presettable 8-Bit down Counter
●40192	Sync Up/Down Decade Counter
●40193	Sync Up/Down Binary Counter
●40160	Sync Decade Counter
●40161	Sync Binary Counter
●40162	Fully Sync Decade Counter
●40163	Fully Sync Binary Counter
74C90	Decade Counter
74C93	4-Bit Binary Counter
74C160	Sync Decade Counter
74C161	Sync Binary Counter
74C162	Fully Sync Decade Counter
74C163	Fully Sync Binary Counter
74C192	Sync Up/Down Decade Counter
74C193	Sync Up/Down Binary Counter
74C925	4 Decade Counter Divider
74C926	4 Decade Counter Divider with Display Select
74C927	Minutes and Seconds Counter Driver
74C928	2-1/2 Decade Counter Driver

Display drivers

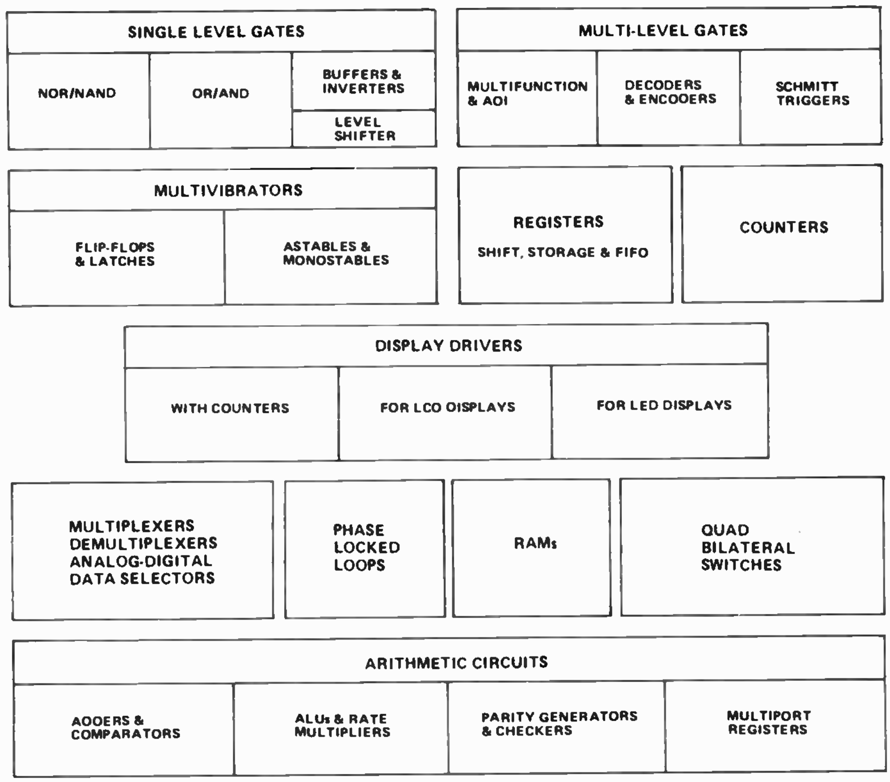
with counters

4026	Decade Counter/Divider with Seven-Segment Display outputs and Display Enable.
4033	Decade Counter/Divider with 7-Segment Display Outputs and Ripple Blanking.

for LCD displays

4054	4-Line Liquid Crystal display Driver.
4055	BCD-to-Seven-Segment Decoder/Driver with 'Display-Frequency' Output, for Liquid Crystal Displays.

CMOS



TTL to CMOS. Functionally Equivalent Types

TTL	CMOS		TTL	CMOS		TTL	CMOS	
7400	4011		7475	4042		74150	4067	
7401	40107		7476	4027		74151	4051	4097
7402	4001		7477	4042		74152	4051	4097
7404	4009	4049	7478	4027		74153	4052	
7406	4009	4049	7483	4008		74154	4514	4515
7407	4010	4050	7485	4063		74155	4555	4556
7408	4081		7486	4030	4070	74156	4555	4556
7410	4023		7490	4518		74157	4019	
7411	4073		7491	4015	4094	74164	4015	
7420	4012		7493	4520		74165	4021	
7425	4002		7494	4035		74166	4014	
7427	4025		7495	40104	40194	74167	4527	
7428	4001		7499	40104	40194	74173	4076	
7430	4068		74100	4034		74178	4035	
7432	4071		74104	4095		74179	4035	
7437	4011		74105	4095		74180	40101	
7440	4012		74107	4027		74181	40181	
7442	4028		74110	4095		74182	40182	
7445	4028		74111	4027		74190	4510	
7446	4511	4055	74121	4047	4098	74191	4516	
7447	4511	4055	74122	4047	4098	74194	40104	40194
7448	4511	4055	74123	4098		74195	4035	
7449	4511	4055	74125	4502		74198	4034	
7450	4085		74126	4502		74200	4061	
7453	4086		74132	4093		74251	4051	4097
7454	4086		74136	4030	4070	74279	4044	
7470	4096		74141	4028		74283	4008	
7472	4095		74145	4028		74290	4518	
7473	4027		74148	4532		74293	4520	
7474	4013							

4056	BCD-to-Seven-Segment Decoder/Driver with Strobed-Latch function, for Liquid Crystal Displays.

LED displays

●4511	BCD-to-7-Segment Latch/ Decoder/Driver

Multiplexers/demultiplexers Analogue/digital data selectors

●4016	Quad Bilateral Switch
●4019	Quad AND/OR Select Gate
●4051	Single 8 Channel Multiplexer
●4052	Differential 4 Channel Multiplexer
●4053	Triple 2 Channel Multiplexer
●4066	Counter Divider and Oscillator
●4067	16-channel Multiplexer/ Demultiplexer
4097	Differential 8-channel Multiplexer/Demultiplexer
●4519	4-Bit AND/OR Selector
●4539	DUAL 4-Input Multiplexer
●4555	Dual 1-of-4 Decoder/Demultiplexer (Outputs High)
●4556	Dual 1-of 4 Decoder/Demultiplexer (Outputs Low)
40257	Quad AND/OR Data Selector with Three-State Outputs
●4512	8-Channel Data Selector
4529	Dual 4-Channel Analog Data Selector
74C150	16 to 1 Multiplexer
74C151	8-Channel Digital Multiplexer
74C157	Quad 2-Input Multiplexer
82C19	TRI-STATE 16 to 1 Multiplexer

Phase-locked loop

●4046	Phase Locked Loop

RAMs

4039	4-word, 8-bit, direct word-line addressing
4061	256-word by one bit static RAM
4036	4-word, 8-Bit, binary-addressing
40108	4x4 multiport register
●4720	256 x 1 BIT RAM (15V)
●4721	256 x 4 BIT RAM (15V)
74C89	64-Bit RAM
74C200	256-Bit 256 x 1 (RAM)
74C910	256-Bit (74 x 4) RAM
74C920	1 k (256 x 4) RAM
74C921	1 k (256 x 4) Common I/O RAM
74C929	1 k 16-Pin RAM
74C930	1 k 18-Pin RAM

Quad bilateral switches

●4016	Quad Bilateral Switch
●4066	Counter Divider and Oscillator

Arithmetic circuits

ALUs and rate multipliers

4057	4-Bit ALU
4089	Binary Rate Multiplier
4527	BCD Rate/Multiplier
40181	4-Bit ALU
40182	Look-ahead carry block

Adders & comparators

●4008	4-Bit Full Adder with Parallel Carry
4032	Triple Serial Adder (+logic)
4038	Triple Serial Adder (—logic)
4063	4-Bit Magnitude Comparator
●4030	Quad EX OR Gate
●4070	Quad Exclusive OR Gate
●4077	Quad Exclusive NOR Gate
74C83	4-Bit Binary Full Adder
74C85	4-Bit Comparator
74C909	Linear Comparator

Parity generators & checkers

40101	9-Bit parity generator and checker

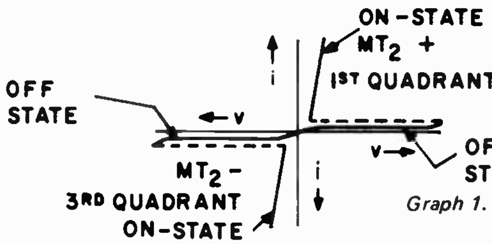
Multiport registers

40108	4x4 multiport register

SC141-146 6A and 10A triacs

The triac is a silicon ac switch which may be gate triggered from an off-state to an on-state for either polarity of applied voltage.

The SC141 and SC146 are molded silicon plastic triacs.



Graph 1. Typical E-I Characteristi

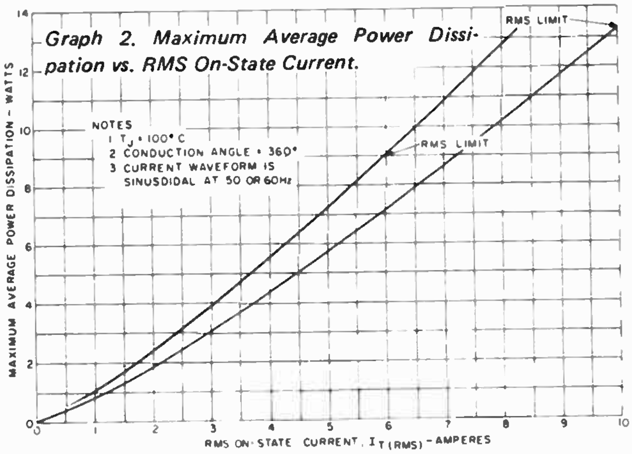
MAXIMUM ALLOWABLE RATINGS

Type	RMS On-State Current, I_T (RMS) Amperes	Repetitive Peak Off-State Voltage, V_{DRM} B Volts	D Volts	E Volts	M Volts	Peak One Full Cycle 50 Hz Surge (Non-Rep) On-State Current, I_{TSM} Amperes	I^2t For Fusing For Times At 1 Millisecond	8.3 Milliseconds
SC141B,D,E,M	6	200	400	500	600	74 A	18 A^2s	26.5 A^2s
SC146B,D,E,M	10	200	400	500	600	90 A	20 A^2s	41.5 A^2s

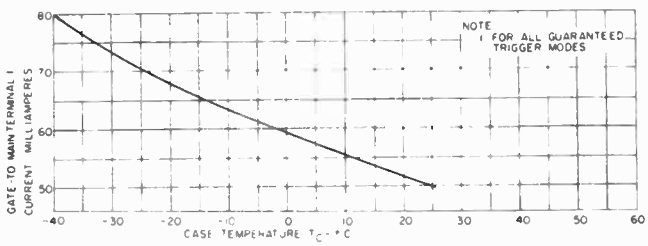
Critical Rate-Of-On-State Current, di/dt:
Breakover voltage triggered operation . . . . 10 A/µs
Peak Gate Power Dissipation, P_{GM} . . . . 10 W for 10 µs
Average Gate Power Dissipation, $P_{G(AV)}$ . . . . 0.5 µs
Peak Gate Current, I_{GM} .see Graph 3
Peak Gate Voltage, V_{GM} .see Graph 3

CHARACTERISTICS – SC 141/SC146

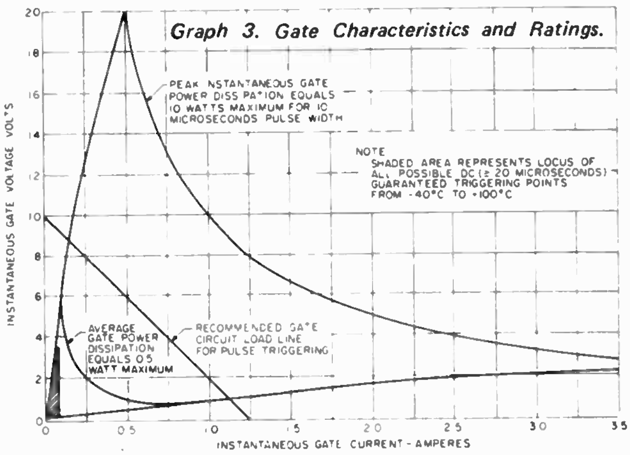
Test	Symbol			Units
Peak On-State Voltage	V_{TM}			
SC141		max.	1.83	Volts
SC146		max.	1.65	
Critical Rate of Rise of Off-State Voltage (Higher values may cause device switching)	dv/dt	typ.	50	Volts/µs
Critical Rate of Rise of Commutating Off-State Voltage (Commutating dv/dt)	$dv/dt_{(c)}$			Volts/µs
		min.	4	
D.C. Gate Trigger Current	I_{GT}	max.	50	mAdc
D.C. Gate Trigger Voltage	V_{GT}	max.	2.5	Vdc
		min.	0.20	
Holding Current	I_H	max.	50	mAdc
Latching Current	I_H			mAdc
MT2+ Gate+		max.	100	
MT2– Gate–		max.	100	
MT2+ Gate–		max.	200	
Steady-State Thermal Resistance Junction to Ambient	RθJA	max.	75	°C/watt
Junction to Case				
SC141	RθJC	max.	3.0	
SC146		max.	2.2	



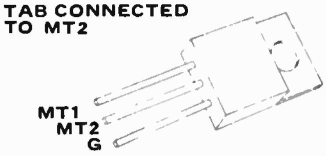
Graph 2. Maximum Average Power Dissipation vs. RMS On-State Current.

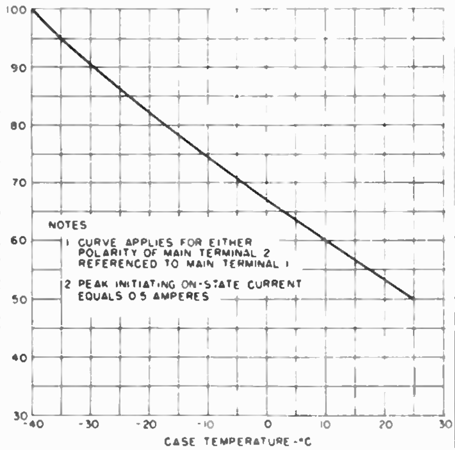


Graph 4. Maximum DC Gate Current to Trigger vs. Case Temperature.



Graph 3. Gate Characteristics and Ratings.

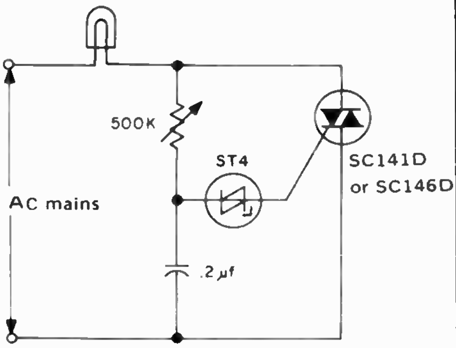




Graph 5. Maximum DC Holding Current vs. Case Temperature.

TYPICAL CIRCUITS

Triacs are especially useful in lamp dimming because of their ability to conduct in both directions.



This circuit uses the ST4 IC to trigger the triac (see p105 for data on this device). The ST4 reduces snap-on effects that occur in symmetrical trigger circuits and minimizes control circuit hysteresis. This performance is possible with a single RC time constant, whereas a symmetrical circuit of comparable performance would require at lease three additional passive components.

ST2 diac

The ST2 diac is a diffused silicon bi-directional trigger diode which may be used to trigger triacs or Silicon Controlled Rectifiers. This device has a three-layer structure having negative resistance switching characteristics for both directions of applied voltage.

MAXIMUM RATINGS ST2

Peak Current (10µsec duration, 120 cycle repetition rate)	I_p	±2 Amperes Max.
Peak Output Voltage	e_p	±3 Volts Min.

CHARACTERISTICS – ST2

Test	Symbols	Min.	Typ.	Max.	Units
Breakover Voltage	$V_{(BR)1}$ and $V_{(BR)2}$	28	32	36	Volts
Breakover Currents	$I_{(BR)1}$ and $I_{(BR)2}$	–	–	200	µamp
Breakover Voltage Symmetry	$V_{(BR)1} - V_{(BR)2}$	–	–	3.8	Volts

ST4 diac

The ST4 is an asymmetrical ac trigger integrated circuit for use in triac phase controls. This device greatly reduces the snap-on effects that are present in symmetrical trigger circuits and minimizes control circuit hysteresis. This performance is possible with a single RC time constant, whereas a symmetrical circuit of comparable performance would require at least three additional passive components.

The ST4 is available in a two-leaded T098 type in-line epoxy package.

ABSOLUTE MAXIMUM RATINGS – ST4

Current

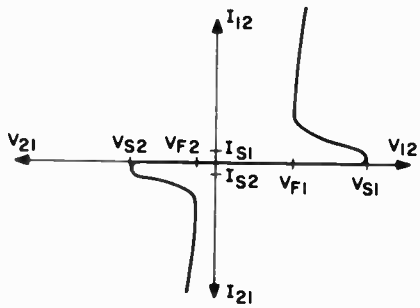
I_{21} Continuous . . . . 200 mA

I_{21} Pulsed (PW = 2µs, Duty Cycle ⩽ 10%) . . . . 500 mA

I_{12} Pulsed (PW = 2µs, Duty Cycle ⩽ 10%) . . . . 175 mA

Power

Total Average . . . . 350 mW



ST4 ELECTRICAL CHARACTERISTICS

ELECTRICAL CHARACTERISTICS

Test	Symbol	Min.	Max.	Units
Switching Voltage	V_{S1}	14	18	V
	V_{S2}	7	9	V
Switching Current	I_{S1}, I_{S2}	–	80	µA
	I_{S1}, I_{S2}	–	160	µA
Voltage Drop	V_{F1}	7	10	V
	V_{F2}	–	1.6	V
Off-State Current	I_{12}	–	100	nA
	I_{21}	–	100	nA
Temperature Coefficient	T.C.	–	.05	%/°C
Turn-on Time	t_{on}	–	1	µs
Turn-off Time	t_{off}	–	30	µs
Output Pulse	V_o	3.5	–	V

SC142 8A triac

The triac is a silicon ac switch which may be gate triggered from an off-state to an on-state for either polarity of applied voltage.

The SC 142 is a molded silicon plastic triac.

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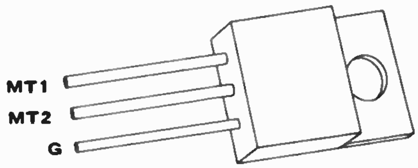
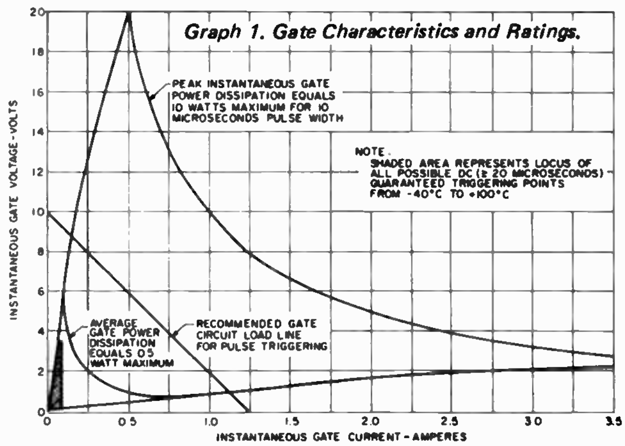


TAB ISOLATED
MT = Main Terminal

MAXIMUM ALLOWABLE RATINGS – SC 142

RMS On-State Current, It (rms)	Repetitive Peak Off-State Voltage VDRM				Peak One Full Cycle (50 Hz) Surge (Non-Rep) On-State Current ITSM Amperes	I^2t for Fusing For Times At	
	SC142B	SC142D	SC142E	SC142M		1 Millisecond	8.3 Millisecond:
8 A	200V	400 V	500 V	600V		18 A^2s	26.5 A^2s

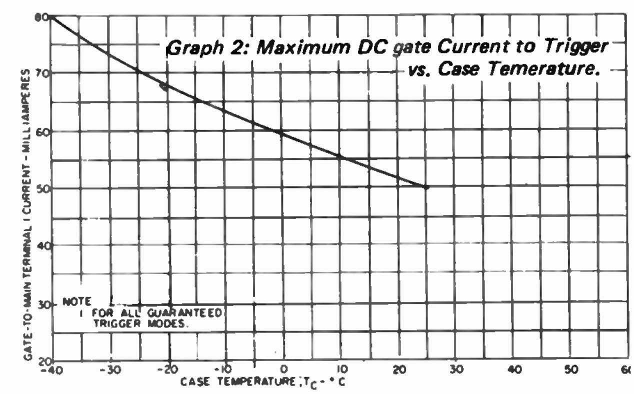
Critical Rate-Of-Rise of On-State Current, di/dt:
Breakover voltage triggered operation . . . . 10 A/µ
Peak Gate Power Dissipation, RGM. . . . . 10 W for 10 µ
Average Gate Power Dissipation, PG(AV) . . . . 0.5 V
Peak Gate Current, IGM . . . . see Graph
Peak Gate Voltage, VGM. . . . . see Graph

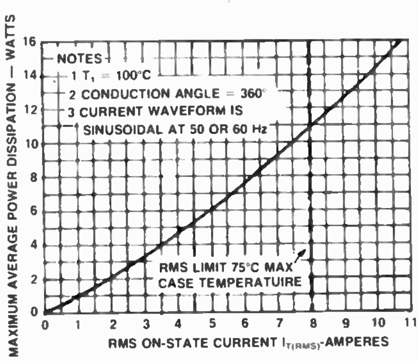


Graph 1. Gate Characteristics and Ratings.

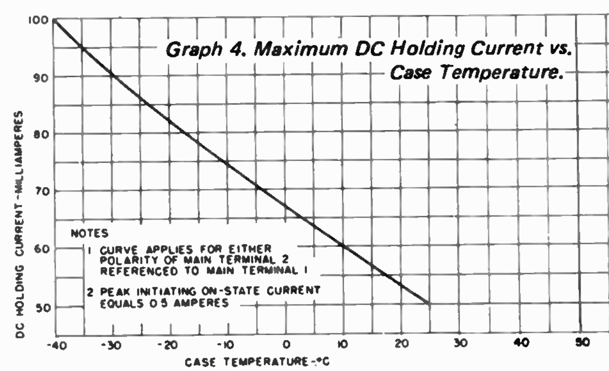
CHARACTERISTICS – SC 142

Test	Symbol			Units
Peak On-State Voltage	V_{TM}	max.	1.75	V
Critical Rate of Rise of Off-State Voltage (Higher values may cause device switching)	dv/dt	typ.	50	V/µs
Critical Rate of Rise of Commutating Off-State Voltage (Commutating dv/dt)	$dv/dt_{(c)}$	min.	4	V/µs
D.C. Gate Trigger Current	I_{GT}	max.	50	mAdc
D.C. Gate Trigger Voltage	V_{GT}	max. min.	2.5 0.20	Vdc
Holding Current	I_H	max.	50	mAdc
Latching Current	I_L			mAdc
MT2+ Gate+		max.	100	
MT2– Gate–		max.	100	
MT2+ Gate–		min.	200	
Steady-State Thermal Resistance				°C/W
Junction to Ambient	$R\theta JA$	max.	75	
Junction to Case	$R\theta JC$	max.	3.3	



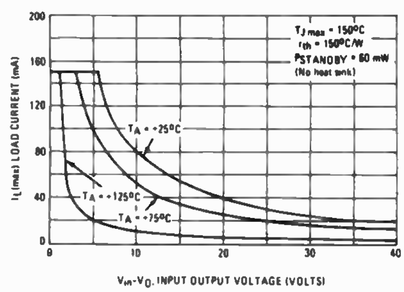


Graph 3. Maximum Average Power Dissipation.

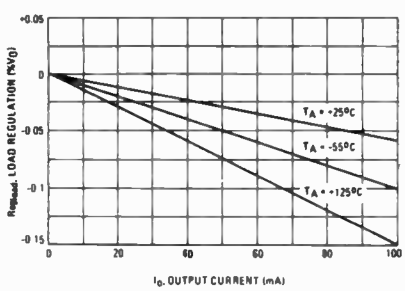


723 universal voltage regulator

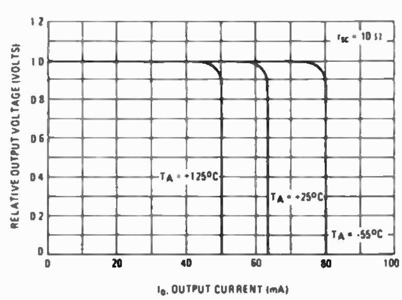
723 Universal Voltage Regulator
The 723 is a positive or negative voltage regulator designed to deliver load currents to 150 mA. The output current can be increased to several amps using one or more external pass transistors. The output voltage is adjustable from 2 to 37 V. Short-circuit protection is adjustable.



Graph 1 Maximum load current as a function of input-output voltage differential.



Graph 2 – Load regulation characteristics without current limiting.

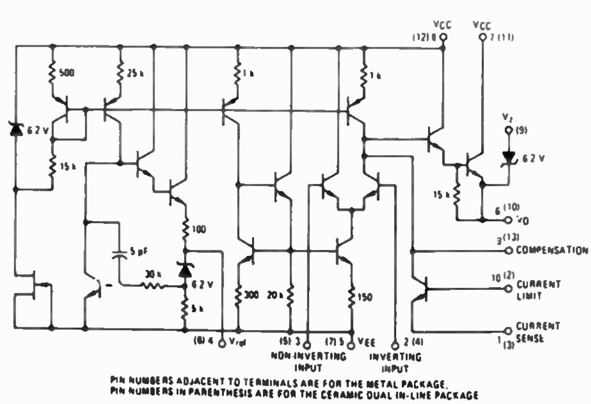


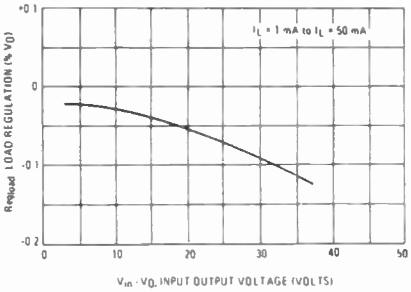
Graph 3 – Current limiting characteristics.

ELECTRICAL CHARACTERISTICS

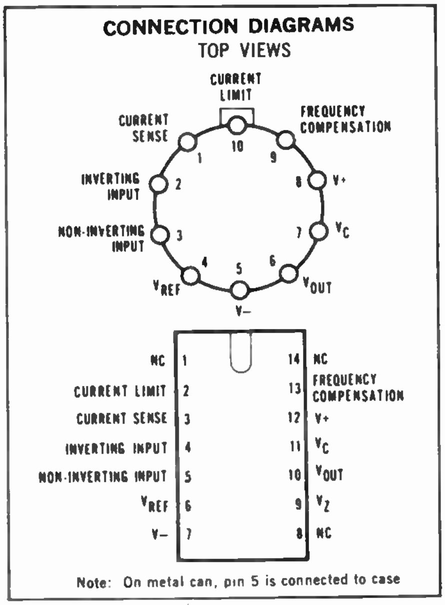
Unless otherwise noted: $T_A = +25°C$, $V_{in} = 12$ Vdc, $V_O = 5$Vdc, $I_L = 1$ mAdc, $r_{sc} = 0$, C1 = 100 pf, $C_{ref} = 0$ and divider impedance as seen by the error amplifier $\leqslant 10\ k\Omega$ connected as shown in Figure 1, $T_{low} = 0°C$, $T_{high} = +75°C$.

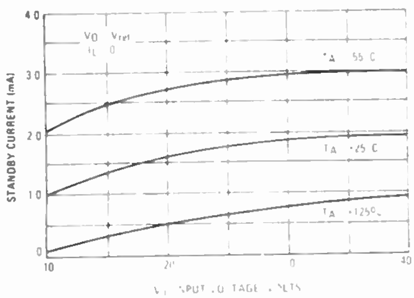
Characteristics	Symbol*	MC1723C Min	Typ	Max	Unit
Input Voltage Range	V_{in}	9.5	–	40	Vdc
Output Voltage Range	V_O	2.0	–	37	Vdc
Reference Voltage	V_{ref}	6.80	7.15	7.50	Vdc
Average Temperature Coefficient of Output Voltage ($T_{low} < T_A < T_{high}$)	TCV_O	–	0.003	0.015	%/°C
Line Regulation	Reg_{in}				%V_O
($T_A = +25°C$) $12\ V < V_{in} < 15\ V$		–	0.01	0.1	
$12\ V < V_{in} < 40\ V$		–	0.1	0.5	
($T_{low} < T_A < T_{high}$) $12\ V < V_{in} < 15\ V$		–	–	0.3	
Load Regulation (1.0 mA L < 50 ma)	Reg_{load}				%V_O
$T_A = +25°C$		–	0.03	0.2	
$T_{low} < T_A < T_{high}$		–	–	0.6	
Ripple Rejection (f = 50 Hz to 10 kHz)	Rej_R				dB
$C_{ref} = 0$		–	74	–	
$C_{ref} = 5.0\ \mu F$		–	86		
Short Circuit Current Limit ($r_{sc} = 10\ \Omega$, $V_O = 0$)	I_{sc}	–	65	–	,Adc



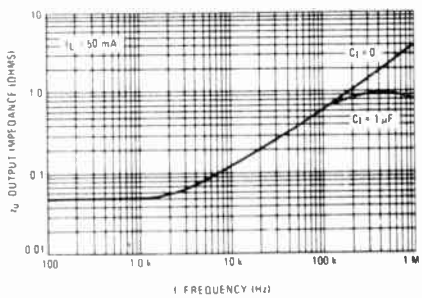


Graph 4 Load regulation as a function of input-output voltage differential.





Graph 5 Standby current drain as a function of input voltage.



Graph 6 – Output impedance as function of frequency.

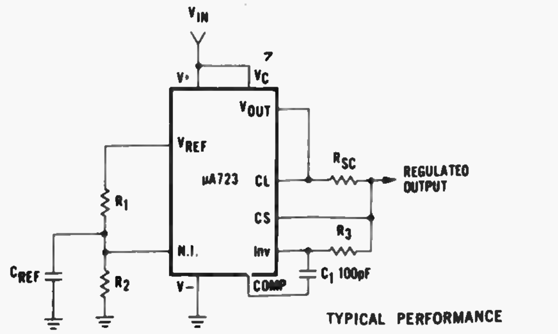
RESISTOR VALUES (kΩ) FOR POSITIVE OUTPUT VOLTAGES

POSITIVE OUTPUT VOLTAGE	APPLICABLE FIGURES	FIXED OUTPUT ±5 percent		OUTPUT ADJUSTABLE ±10 percent		
		R_1	R_2	R_1	P_1	R_2
+3.0	1, 5, 6, 7, (4)	4.12	3.01	1.8	0.5	1.2
+3.6	1, 5, 6, 7, (4)	3.57	3.65	1.5	0.5	1.5
+5.0	1, 5, 6, 7, (4)	2.15	4.99	.75	0.5	2.2
+6.0	1, 5, 6, 7, (4)	1.15	6.04	0.5	0.5	2.7
+9.0	2, 4, (5, 6, 7)	1.87	7.15	.75	1.0	2.7
+12	2, 4, (5, 6, 7)	4.87	7.15	2.0	1.0	3.0
+15	2, 4, (5, 6, 7)	7.87	7.15	3.3	1.0	3.0
+28	2, 4, (5, 6, 7)	21.0	7.15	5.6	1.0	2.0

723 universal voltage regulator

RESISTOR VALUES (k Ω) FOR NEGATIVE OUTPUT VOLTAGES.					
NEGATIVE OUTPUT VOLTAGE	FIXED OUTPUT ±5%		5% OUTPUT ADJUSTABLE ±10%		
	R_1	R_2	R_1	P_1	R_2
−6	3.57	2.43	1.2	0.5	.75
−9	3.48	5.36	1.2	0.5	2.0
−12	3.57	8.45	1.2	0.5	3.3
−15	3.65	11.5	1.2	0.5	4.3
−28	3.57	24.3	1.2	0.5	10

See Figs. 3 and 8 for negative voltage regulation circuits

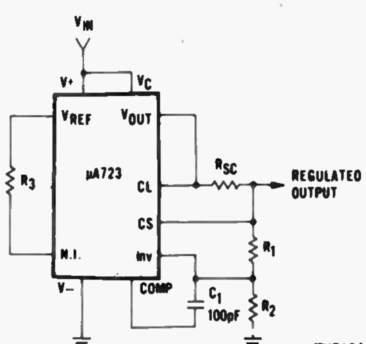


TYPICAL PERFORMANCE

Regulated Output Voltage	5 V
Line Regulation (ΔV_{IN} = 3 V)	0.5 mV
Load Regulation (ΔI_L=50 mA)	1.5 mV

Note: $R_3 = \frac{R_1 R_2}{R_1 + R_2}$ for minimum temperature drift.

Fig.1. Basic low voltage regulator (V_{out} = 2 to 7 volts)



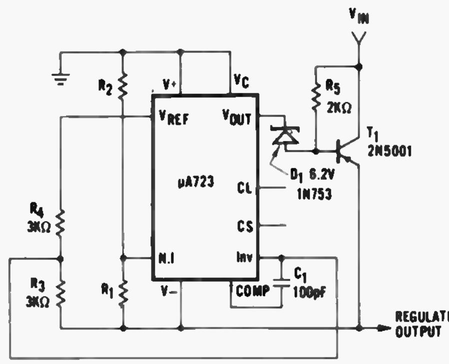
TYPICAL PERFORMANCE

Regulated Output Voltage	15 V
Line Regulation (ΔV_{IN} = 3 V)	1.5 mV
Load Regulation (ΔI_L=50 mA)	4.5 mV

Note: $R_3 = \frac{R_1 R_2}{R_1 + R_2}$ for minimum temperature drift.

R_3 may be eliminated for minimum component count.

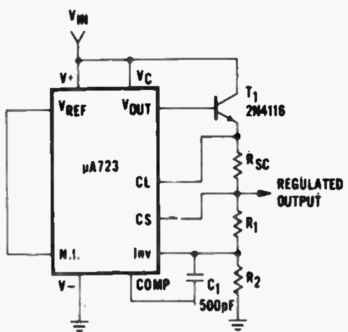
Fig. 2. Basic high voltage regulator (V_{out} = 7 to 37 Volts).



TYPICAL PERFORMANCE

Regulated Output Voltage	−15 V
Line Regulation (ΔV_{IN} = 3 V)	1 mV
Load Regulation (ΔI_L = 100 mA)	2 mV

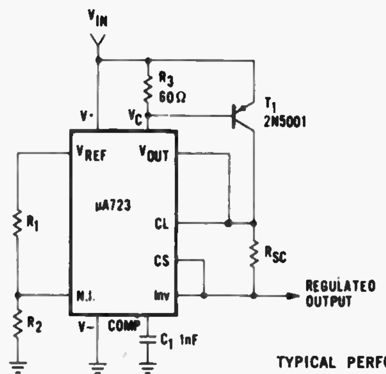
Fig. 3. Negative voltage regulator.



TYPICAL PERFORMANCE

Regulated Output Voltage	+15 V
Line Regulation (ΔV_{IN} = 3 V)	1.5 mV
Load Regulation (ΔI_L = 1 A)	15 mV

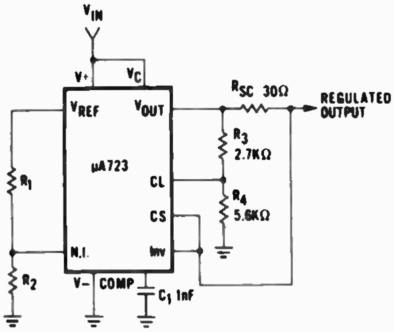
Fig. 4. Positive voltage regulator (External NPN Pass Transistor).



TYPICAL PERFORMANCE

Regulated Output Voltage	+5 V
Line Regulation (ΔV_{IN} = 3 V)	0.5 mV
Load Regulation (ΔI_L = 1 A)	5 mV

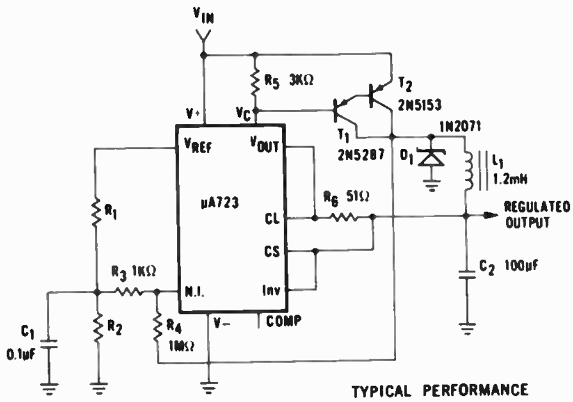
Fig. 5. Positive voltage regulator (External PNP Pass Transistor).



TYPICAL PERFORMANCE

Regulated Output Voltage	+5 V
Line Regulation ($\Delta V_{IN} = 3$ V)	0.5 mV
Load Regulation ($\Delta I_L = 10$ mA)	1 mV
Current Limit Knee	20 mA

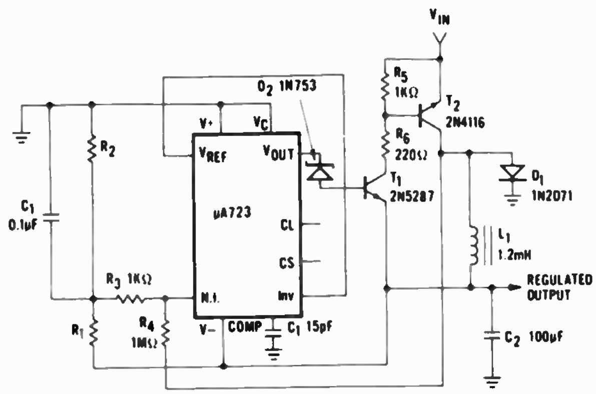
Fig. 6. Foldback current limiting.



TYPICAL PERFORMANCE

Regulated Output Voltage	+5 V
Line Regulation ($\Delta V_{IN} = 30$ V)	10 mV
Load Regulation ($\Delta I_L = 2$ A)	80 mV

Fig. 7. Positive switching regulator.



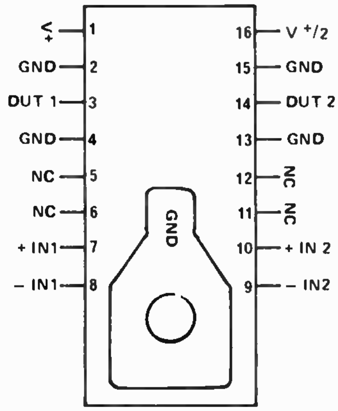
TYPICAL PERFORMANCE

Regulated Output Voltage	−15 V
Line Regulation ($\Delta V_{IN} = 20$ V)	8 mV
Load Regulation ($\Delta I_L = 2$ A)	6 mV

Fig. 8. Negative switching regulator.

LM379 dual 6 watt audio amplifier

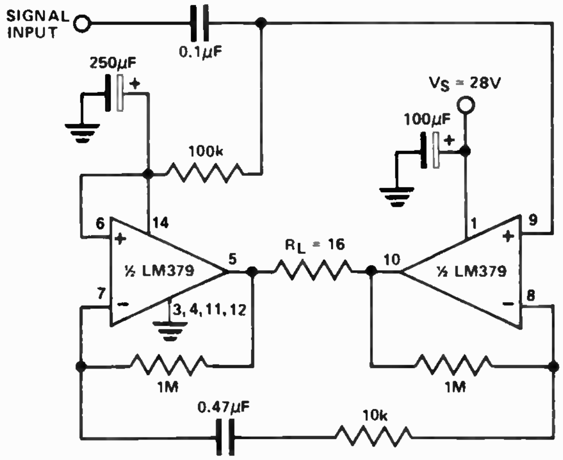
The LM379 is a monolithic dual power amplifier which offers high quality performance. The amplifier is designed to operate with a minimum of external components and contains an internal bias regulator to bias each amplifier. Overload protection consisting of both internal current limit and thermal shutdown is provided.



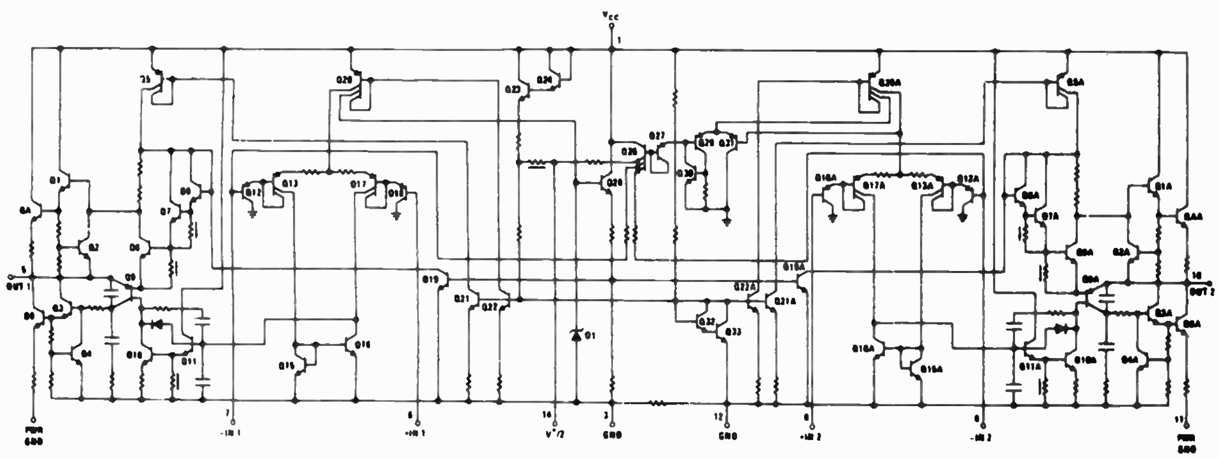
QUICK REFERENCE DATA

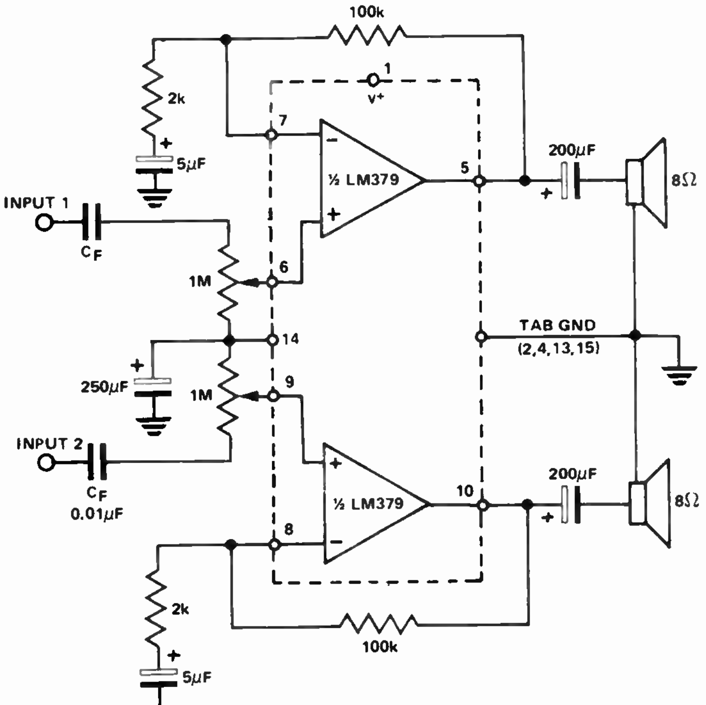
Absolute Maximum Ratings

Supply voltage		35 V
Input voltage		0 V – V_{SUPPLY}
Operating temperature		0°C to 70°C
Characteristics		
With 28 V supply, 8Ω load, and 34 dB gain.		
Supply current (P_{OUT} = 0 W)	typ.	15 mA
Supply current (P_{OUT} = 1.5 W/channel)	typ.	430 mA
DC output level	typ.	14 V
Supply voltage	min.	10 V
Output power (THD = 5%)	typ.	6 W
Output power (THD = 10%)	typ.	7 W
Total harmonic distortion (1 kHz, 1 W)	typ.	0.07%
Total harmonic distortion (1 kHz, 4 W)	typ.	0.2%
Offset voltage	typ.	15 mV
Input bias current	typ.	100 nA
Input impedance	min.	3 MΩ
Open loop gain (R_S = 0Ω)	typ.	90 dB
Channel separation (C_F = 250 μF, f=1 kHz)	typ.	70 dB
Ripple rejection (f = 120 Hz, C = 250 μF)	typ.	70 dB
Current limit	typ.	1.5 A
Slew rate	typ.	1.4 V/μs
Equivalent input noise voltage (R = 600Ω, 100 Hz – 10 kHz)	typ.	3 μVrms

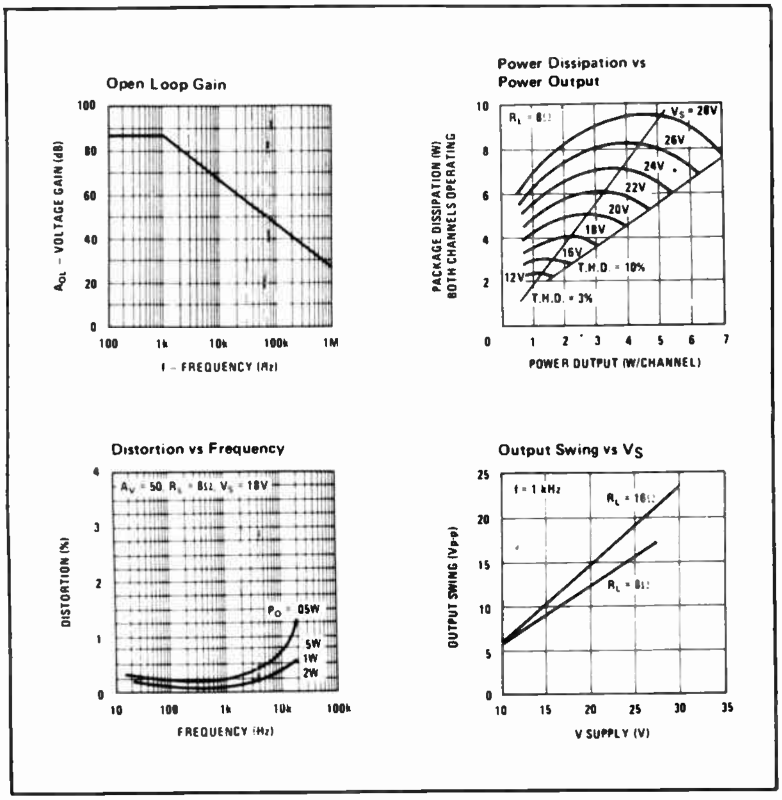


14 W Bridge amplifier.





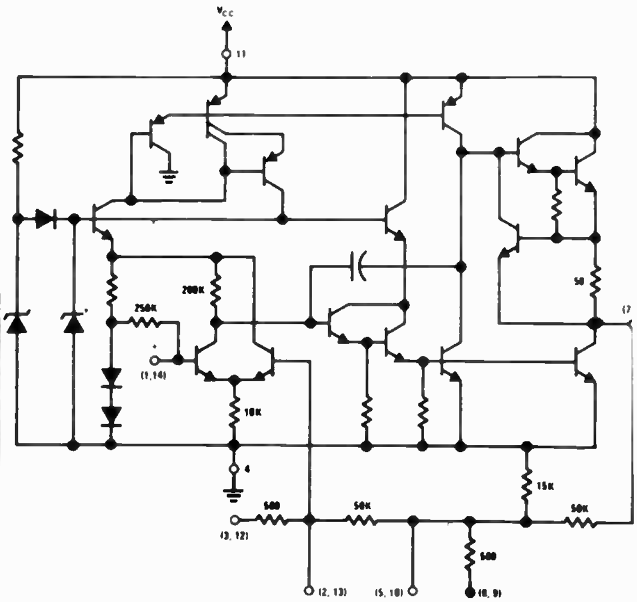
Simple stereo amplifier.



LM382 low noise dual preamplifier

The LM382 is a dual preamplifier for low level applications requiring optimum noise performance. Each amplifier is completely independent, with individual internal power supply decoupler-regulator, providing 120 dB supply rejection and 60 dB channel separation. Other features include high gain (100 dB), large output voltage swing (V_{cc} -2 V), and wide power bandwidth (75 kHz, 20 V_{cc}). The LM382 operates from a single supply across the wide range of 9 to 40 V.

A resistor matrix is provided on the chip to allow the user to select a variety of closed loop gain options and frequency response characteristics such as flat-band. NAB or RIAA equalization.



QUICK REFERENCE DATA

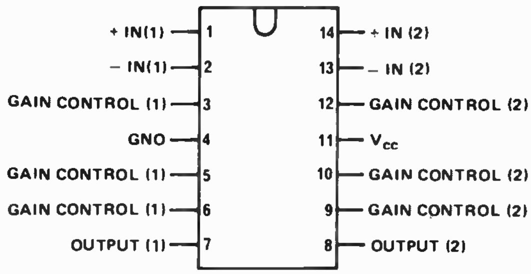
Absolute Maximum Ratings

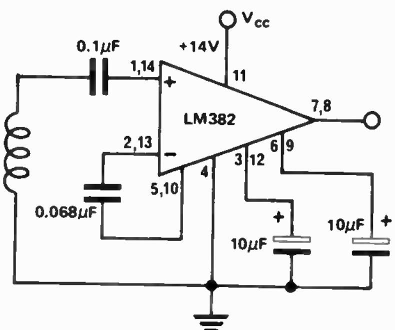
Supply voltage	max.	+ 40 V
Power dissipation	max.	800 mW
Operating temperature	range	0°C to 70°C

Characteristics

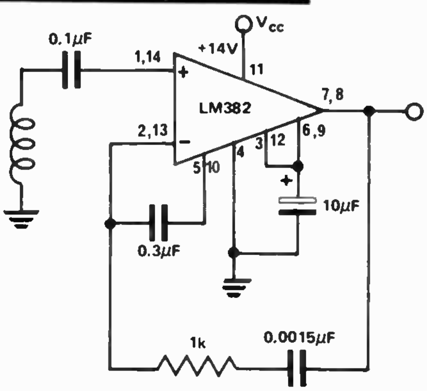
with 14 V supply

Voltage gain (Open loop, differential input)	typ.	100 000 V/V
Supply current ($R_L = \infty$)	typ.	10 mA
Input resistance (positive input)	typ.	100 kΩ
Input resistance (negative input)	typ.	200 kΩ
Input current (negative input)	typ.	0.5 μA
Output resistance (open loop)	typ.	150 Ω
Output current (source)	typ.	8 mA
Output current (sink)	typ.	2 mA
Output voltage swing (pk-pk, R=10 k)	typ.	V_{CC} – V
Small signal bandwidth	typ.	15 MHz
Power bandwidth (20 $V_{p\text{-}p}$ V = 24 V)	typ.	75 kHz
Maximum input voltage (linear operation)	max.	300 mVrms
Supply rejection ratio (f=1 kHz)	typ.	120 dB
Channel separation (f=1 kHz)	typ.	60 dB
Total harmonic distortion (60 dB gain, f=1 kHz)	typ.	0.1%
Total equivalent input noise (R=600, 100 – 10 000 Hz)	typ.	0.8 Vrms
Noise figure, 100 – 10 000 Hz (50 kΩ)	typ.	1.0 dB
Noise figure, 100 – 10 000 Hz (10 kΩ)	typ.	1.6 dB
Noise figure, 100 – 10 000 Hz (5 kΩ)	typ.	2.8 dB

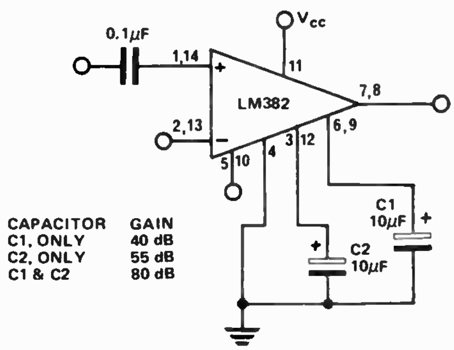




Tape Preamplifier (NAB equalisation).

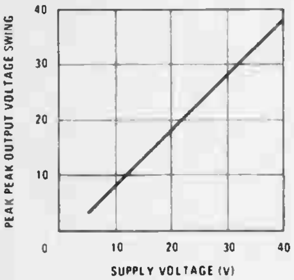


Phono pre-amp (RIAA equalisation).

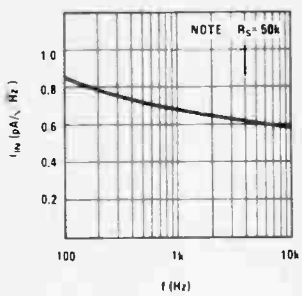


Flat response – Fixed gain configuration.

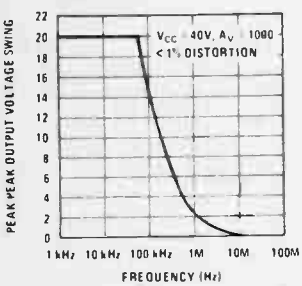
P–P Output Voltage Swing vs V_{CC}



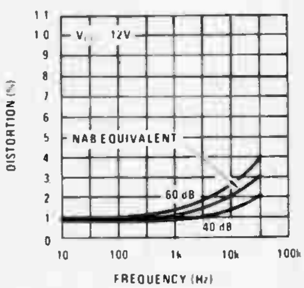
Noise Current vs Frequency



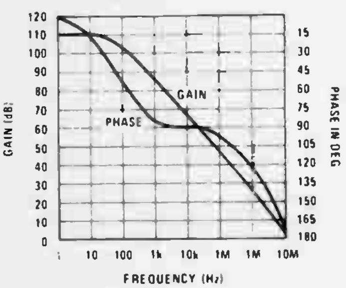
Large Signal Frequency Response



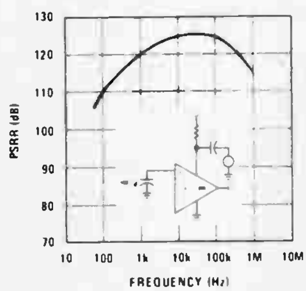
% Distortion



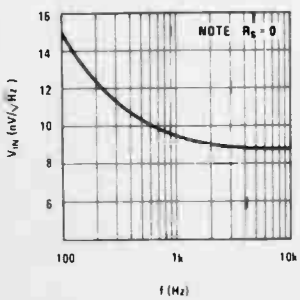
Gain and Phase Response



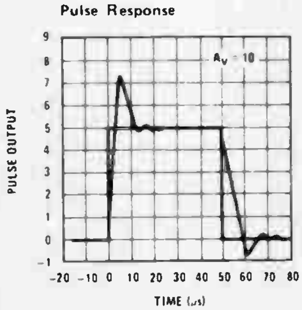
PSRR vs Frequency



Noise Voltage vs Frequency



Pulse Response

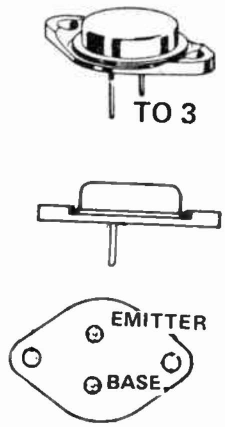


2N3055 npn silicon power transistor

The 2N3055 and it usual pnp complement, MJ2955, are popular high-power transistors in metal packages. The 2955 is dealt with in an accompanying data sheet.

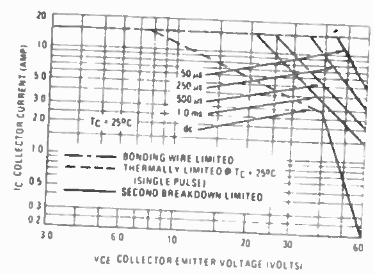
Motorola and Texas Instruments manufacture plastic encapsulated versions of this pair with the following type numbers: MJE3055/MJE2955, and TIP3055/TIP2955, respectively. The plastic types cannot handle the same collector currents as the metal types; see the graphs of the safe operating areas. The total device power dissipation of the pnp types is: 115 W (2N3055), and 90 W (MJE3055 and TIP3055). Recent ETI projects using this transistor include: — ETI 240 Emergency flash, — ETI 541 Train controller, — ETI 780 Novice transmitter, — ETI 131 General purpose power supply.

2N3055 package



COLLECTOR connected to mounting base

2N3055 safe operating area



QUICK REFERENCE DATA 2N3055

Maximum ratings

Collector-emitter voltage (V_{CEO})	60 V
Collector-emitter voltage (V_{CER})	70 V
Collector-base voltage (V_{CB})	100 V
Emitter-base voltage (V_{EB})	7 V
Collector current (continuous) (I_C)	15 A
Power dissipation (P_D)	115 W

Off characteristics

Collector-emitter sustaining voltage (I_C = 0.2 A, I_B = 0)	V_{CEO} (sus)	min. 60 V
Collector-emitter sustaining voltage (I_C = 0.2 A, R_{BE} = 100 ohms)	V_{CER} (sus)	min. 70 V
Collector cut-off current (V_{CE} = 30 V, I_B = 0)	I_{CEO}	max. 0.7 mA
Emitter cut-off current (V_{BE} = 7 V, I_C = 0)	I_{EBO}	max. 5 mA

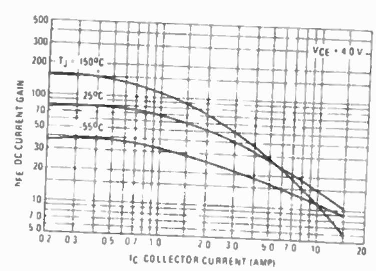
On Characteristics

DC current gain	h_{FE}	
(I_C = 4 A, V_{CE} = 4 V)		min. 20; max. 70
(I_C = 10 A, V_{CE} = 4 V)		min. 5
Collector-emitter saturation voltage	V_{CE} (sat)	
(I_C = 4 A, I_B = 0.4 A)		max. 1.1 V
(I_C = 10 A, I_B = 3.3 A)		max. 3 V
Base-emitter on voltage (I_C = 4 A, V_{CE} = 4 V)	V_{BE} (on)	max. 1.5 V
Second breakdown collector current with base forward biased (t ; 1 s, V_{CE} = 40 V)	$I_{S/b}$	min. 2.87 A

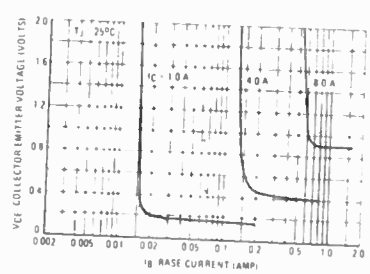
Dynamic Characteristics

Current gain – bandwidth product (I_C = 1 A, V_{CE} = 4 V, f_{test} = 1 MHz)	f_T	min. 2.5 MHz
Small-signal current gain (I_C = 1A, V_{CE} = 4 V, f = 1 kHz)	h_{fe}	min. 15 max. 120

2N3055 DC current gain



2N3055 Collector saturation region



Safe operation

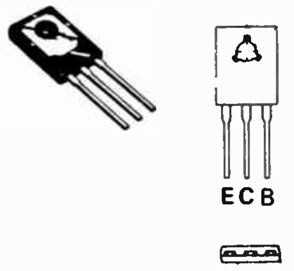
For the three types we have mentioned in this data sheet we have printed the most important graph of a transistor's operating characteristics: the active-region safe operating area graph. These graphs show the different limits on use of the plastic and metal types, and the experimenter ought to be familiar with their interpretation.

There are two limitations on the power handling ability of a transistor: average junction temperature and second breakdown. Safe operating area curves indicate IC — VCE limits of the transistor that must be observed for reliable operation, i.e., the transistor must not be subjected to greater dissipation than the curves indicate.

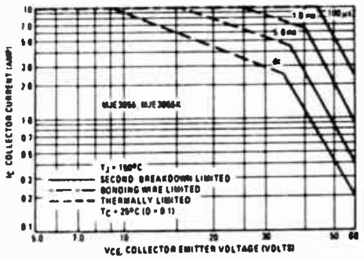
Second breakdown pulse limits are valid for duty cycles to10%.At high case temperatures, thermal limitations may reduce the power that can be handled to values less than the limitations imposed by second breakdown.

NOTE Beware using transistors with these type numbers followed by suffix 1 or suffix 2. These are low voltage rejects and it is important to establish their limits before using them in power circuits.

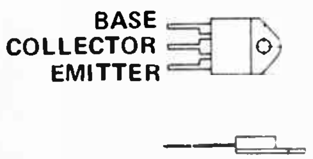
MJE3055



MJE3055 safe operating area.

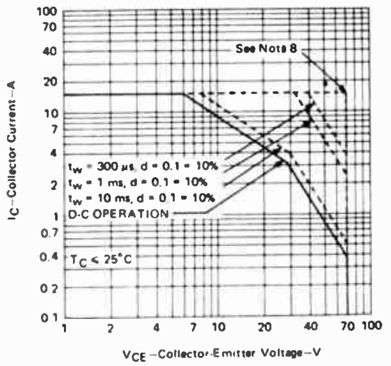


TIP3055



The collector is in electrical contact with the mounting tab

TIP3055 safe operating area



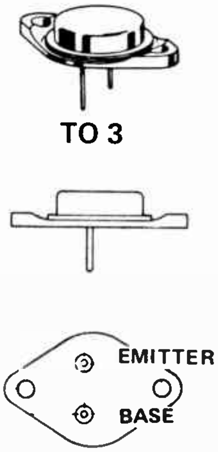
MJ2955 pnp silicon power transistor

The MJ2955 is the usual pnp complement of the 2N3055 but there are some differences, the most obvious being the higher thermal limit which means the maximum power dissipation of the device is 150 W (compared to 115 W for the 2N3055).

The MJ2955 is in a metal package but plastic types are available: MJE2955 (90 W) and TIP2955 (90 W). The safe operating area graphs should be studied before using these types.

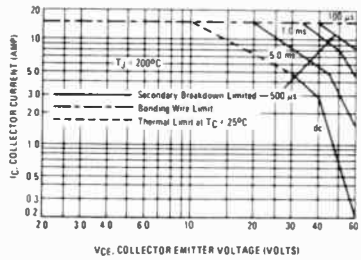
This transistor has been used in quite a few ETI projects: ETI541 Train controller, ETI Swimming pool alarm, ETI422 Fifty watt amplifier.

MJ2955 package

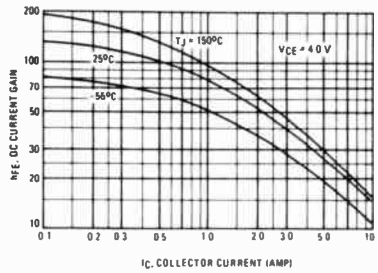


COLLECTOR connected to mounting base

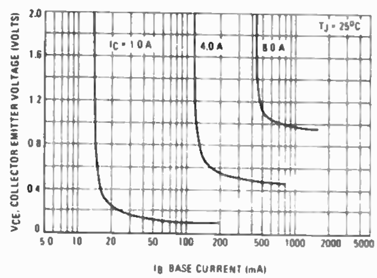
MJ2955 safe operating area



MJ2955 DC current gain



MJ2955 collector saturation region



MJ2955 QUICK REFERENCE DATA

Maximum ratings

Collector-emitter voltage (V_{CEO})	60 V
Collector-emitter voltage (V_{CER})	70 V
Collector-base voltage (V_{CB})	100 V
Emitter-base voltage (V_{EB})	7 V
Collector current (continuous) (I_C)	15 A
Base current (I_B)	7 A
Power dissipation (P_D)	150 W

Off Characteristics

Collector-emitter sustaining voltage (I_C = 0.2 A, I_B = 0)	V_{CEO} (sus)	min. 60 V
Collector-emitter breakdown voltage (I_C = 0.2 A, R_{BE} = 100 ohms)	BV_{CER}	min. 70 V
Collector cut-off current (V_{CE} = 30 V, I_B = 0)	I_{CEO}	max. 0.7 mA
Emitter cut-off current (V_{BE} = 7 V, I_C = 0)	I_{EBO}	max. 5 mA

On Characteristics

DC current gain	h_{FE}	
(I_C = 4 A, V_{CE} = 4 V)		min. 20; max. 70
(I_C = 10 A, V_{CE} = 4 V)		min. 5
Collector-emitter saturation voltage	V_{CE} (sat)	
(I_C = 4 A, I_B = 0.4 A)		max. 1.1 V
(I_C = 10 A, I_B = 3.3 A)		max. 3 V
Base-emitter on voltage (I_C = 4 A, V_{CE} = 4 V)	V_{BE} (on)	max. 1.8 V

Dynamic Characteristics

Current gain – bandwidth product (I_C = 0.5 A, V_{CE} = 10 V, f_{test} = 1 MHz)	f_T	min. 4 MHz
Small-signal current gain (I_C = 1A, V_{CE} = 4 V, f = 1 kHz)	h_{fe}	min. 15
Small-signal current gain cut-off frequency (V_{CE} = 4V, I_C = 1 A, f = 1 kHz)	$f_{\alpha e}$	min. 10 kHz

Notes

Safe operation

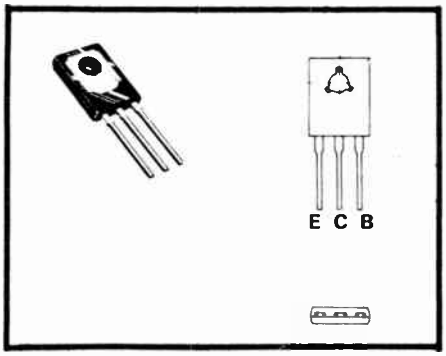
For the three types we have mentioned in this data sheet we have printed the most important graph of a transistor's operating characteristics: the active-region safe operating area graph. These graphs show the different limits on use of the plastic and metal types, and the experimenter ought to be familiar with their interpretation.

There are two limitations on the power handling ability of a transistor: average junction temperature and second breakdown. Safe operating area curves indicate IC — VCE limits of the transistor that must be observed for reliable operation, i.e., the transistor must not be subjected to greater dissipation than the curves indicate.

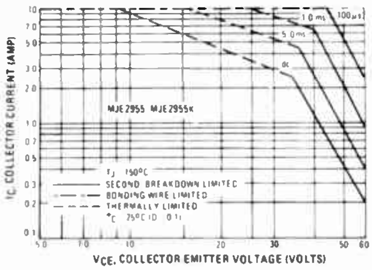
Second breakdown pulse limits are valid for duty cycles to10%.At high case temperatures, thermal limitations may reduce the power that can be handled to values less than the limitations imposed by second breakdown.

NOTE Beware using transistors with these type numbers followed by suffix 1 or suffix 2. These are low voltage rejects and it is important to establish their limits before using them in power circuits.

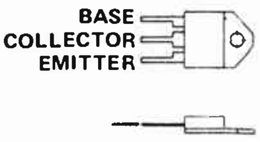
MJE2955



MJE2955 safe operating area

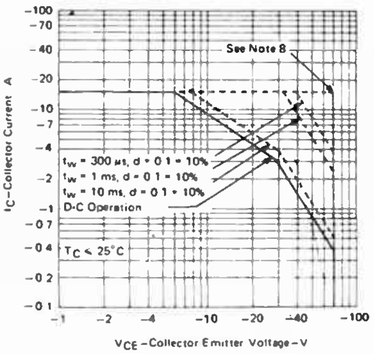


TIP2955



The collector is in electrical contact with the mounting tab.

TIP2955 safe operating area.



BC107-109 low power amplifiers

QUICK REFERENCE DATA

			BC107	BC108	BC109
Collector-emitter voltage (V_{BE} = 0)	V_{CES}	max.	50	30	30 V
Collector-emitter voltage (open base)	V_{CEO}	max.	45	20	20 V
Collector current (peak value)	I_{CM}	max.	200	200	200 mA
Total power dissipation up to T_{amb} = 25°C	P_{tot}	max.	300	300	300 mW
Junction temperature	T_j	max.	175	175	175 °C
Small signal current gain at T_j = 25°C; I_C = 2 mA; V_{CE} = 5 V; f = 1 kHz	h_{fe}	>	125	125	240
		<	500	900	900
Transition frequency at f = 35 MHz; I_C = 10 mA; V_{CE} = 5V	f_T	typ.	300	300	300 MHz
Noise figure at R_S = 2 kΩ; I_C = 200 µA; V_{CE} = 5 V; f = 30 Hz to 15 kHz	F	typ.			1.4 dB
		<			4 dB
f = 1 kHz; Bw = 200 Hz	F	typ.	2	2	1.2 dB
Collector current (dc)	$I_{C\,max}$				100 mA
Thermal Resistance					
From junction to ambient in free air	$R_{th\,j\text{-}a}$	=			0.5 °C/mW
From junction to case	$R_{th\,j\text{-}c}$	=			0.2 °C/mW
Saturation voltages 1); I_C = 10 mA; I_B = 0.5 mA	V_{CEsat}	typ.		90 mV	
		<		250 mV	
	V_{BEsat}	typ.		700 mV	
I_C = 100 mA; I_B = 5 mA	V_{CEsat}	typ.		200 mV	
		<		600 mV	
	V_{BEsat}	typ.		900 mV	
Collector capacitance at f = 1 MHz; I_E = I_e = 0; V_{CB} = 10V	C_c	typ.		2.5 pF	
		<		4.5 pF	
Emitter capacitance at f = 1 MHz; I_C = I_c = 0; V_{EB} = 0.5 V	C_e	typ.		9 pF	

V_{BEsat} decreases by about 1.7 mV/°C with increasing temperature

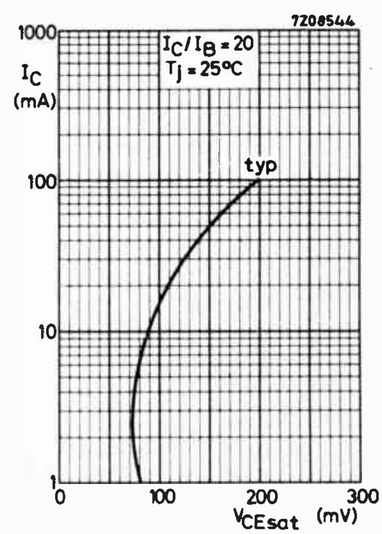
			BC107A BC108A	BC107B BC108B BC109B	BC108C BC109C
DC current gain; I_C = 10 µA; V_{CE} = 5 V	h_{FE}	>		40	100
		typ.	90	150	270
I_C = 2 mA; V_{CE} = 5 V	h_{FE}	>	110	200	420
		typ.	180	290	520
		<	220	450	800

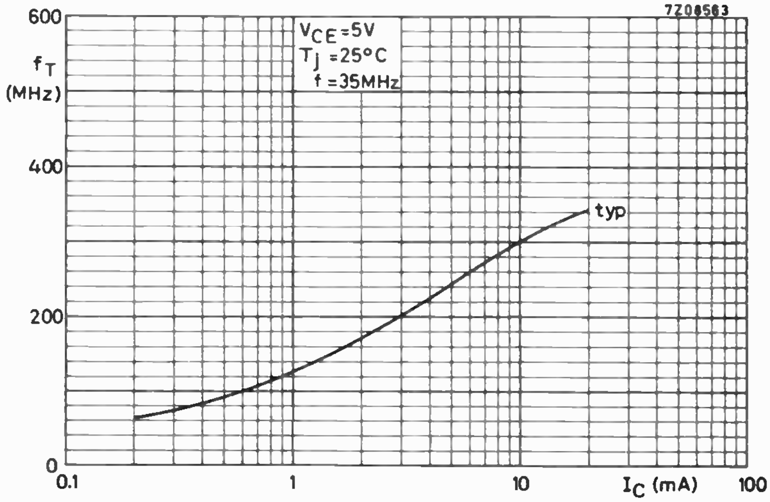
The BC 107-109 series are npn low power transistors commonly used in audio frequency applications. They are housed in TO-18 metal cases with the collector connected to the case. Their pnp complement is the BC 177-179.

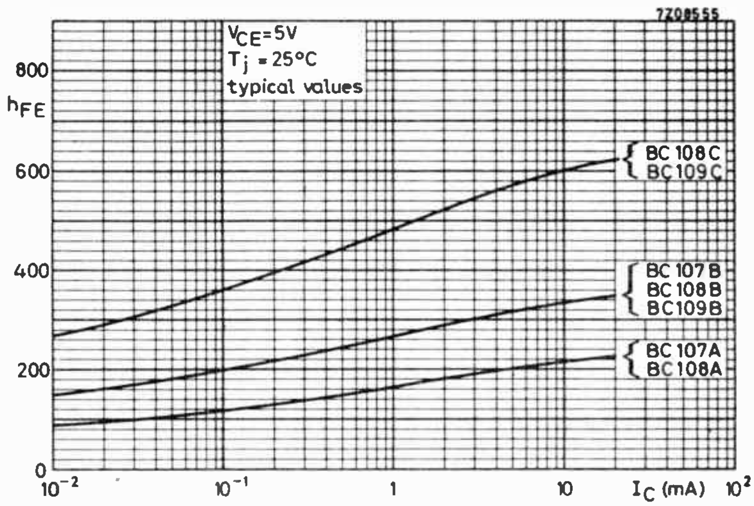
The BC107 is primarily intended for use in driver stages of audio amplifiers and in signal processing circuits of television receivers.

The BC108 is suitable for a multitude of low voltage applications e.g. driver stages of audio pre-amplifiers, and in signal processing circuits of television receivers. The BC109 is primarily intended for low noise input stages in tape recorders, hi-fi amplifiers and other audio frequency equipment.

Each type is made in forms A, B and C — the main difference being in gain (hFE). The A version has the lowest gain — the C version has the highest. Details are included in the Quick Reference Data panel.

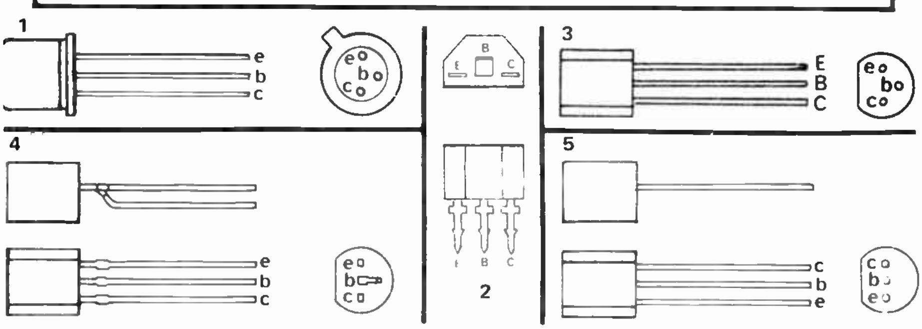






Transistor types				Connections
	BC107	BC108	BC109	1
Electrically similar types	BC147	BC148	BC149	2
	BC237	BC238	BC239	3
	BC547	BC548	BC549	4* 5‡

NOTE: * Philips types only ‡ all others

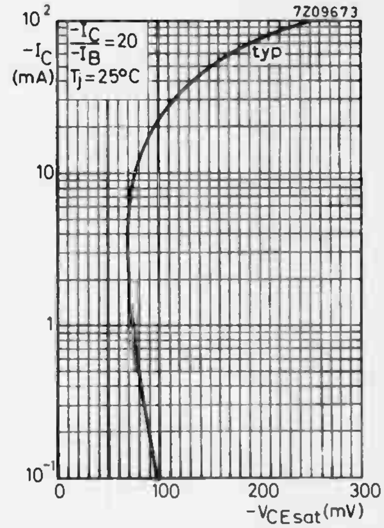


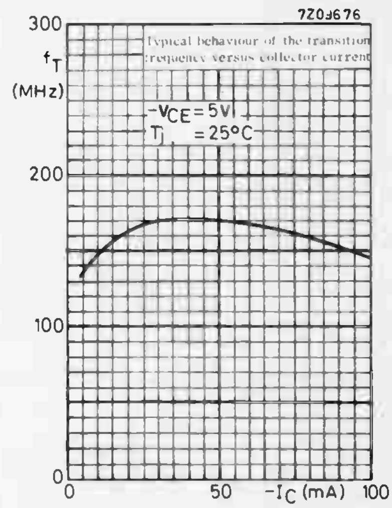
BC177-179 low power amplifiers

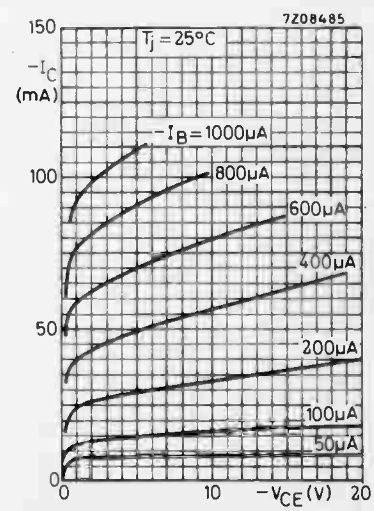
The BC 177-179 series are pnp low power transistors commonly used in audio frequency applications. They are housed in TO-18 metal cases with the collector connected to the case. Their npn complement is the BC 107-109.

The BC177 is a high voltage type primarily intended for use in driver stages of audio amplifiers, and in signal processing circuits of television receivers. The BC178 is suitable for a multitude of low voltage applications e.g. driver stages or audio pre-amplifiers and in signal processing circuits of television receivers. The BC179 is primarily intended for low noise input stages in tape recorders, hi-fi amplifiers and other audio frequency equipment.

As with the BC 107-109 series each BC 177-179 type is manufactured in forms A, B or C — the main difference being in gain (h_{FE}). The A version has the lowest gain — the C version has the highest. Details are included in the Reference Data panel.



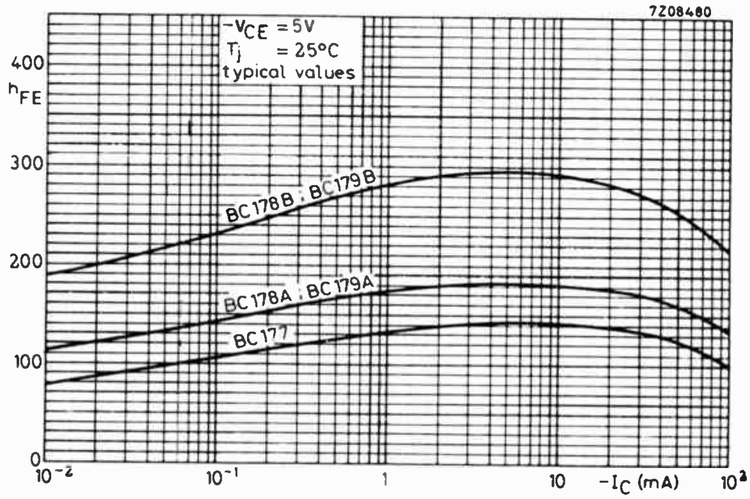




QUICK REFERENCE DATA

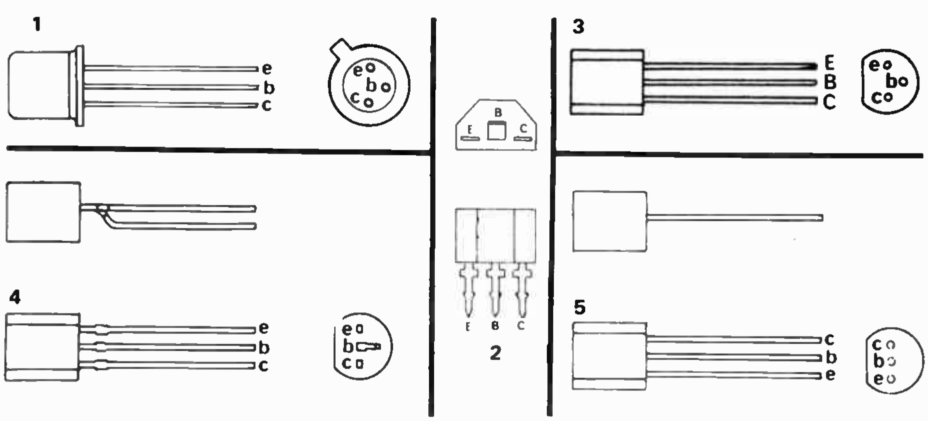
			BC177	BC178	BC179
Collector-emitter voltage ($+V_{BE}$ = 1 V)	$-V_{CEX}$	max.	50	30	25 V
Collector-emitter voltage (open base)	$-V_{CEO}$	max.	45	25	20 V
Collector current (peak value)	$-I_{CM}$	max.	200	200	200 mA
Total power dissipation up to Tamb = 25°C	P_{tot}	max.	300	300	300 mW
Junction temperature	T_j	max.	175	175	175°C
Small signal current gain at Tj = 25°C					
$-I_C$ = 2 mA; $-V_{CE}$ = 5 V; f = 1 kHz	h_{fe}	>	75	75	125
		<	260	500	500
Transition frequency at f = 35 MHz					
$A \cdot I_C$ = 10 mA; $-V_{CE}$ = 5V	f_T	typ.	150	150	150 MHz
Noise figure at R_S = 2 kΩ					
$-I_C$ = 200 µA; $-V_{CE}$ = 5V					
f = 30 Hz to 15 kHz	F	typ.			1.2 dB
		<			4 dB
f = 1 kHz; Bw = 200 Hz	F	<	10	10	4 dB
Collector current (dc)	$-I_C$	max.			100 mA
Thermal Resistance					
From junction to ambient in free air	$R_{th\ j\text{-}a}$	=			0.5 °C/mW
Fron junction to case	$R_{th\ j\text{-}c}$	=			0.2 °C/mW
Saturation voltages					
$-I_C$ = 10 mA; $-I_B$ = 0.5 mA	$-V_{CE\ sat}$	typ.			75 mV
		<			300 mV
$-I_C$ = 100 mA; $-I_B$ = 5 mA	$-V_{BE\ sat}$	typ.			700 mV
	$-V_{CE\ sat}$	typ.			250 mV
	$-V_{BE\ sat}$	typ.			850 mV
Collector capacitance at f = 1 MHz					
$I_E = I_e = 0$; $-V_{CB}$ = 10 V	C_c	typ.			4.0 pF

			BC177	BC178A BC179A	BC178B BC179B
DC current gain					
$-I_C$ = 2 mA; $-V_{CE}$ = 5V	h_{FE}	typ.	140	180	290



Transistor types		Connections
	BC 177 BC 178 BC 179	1
Electrically similar types	BC 157 BC 158 BC 159	2
	BC 307 BC 308 BC 309	3
	BC 557 BC 558 BC 559	4* 5‡

Note: * Philips types only ‡ all others



Notes

BD135-139 general purpose npn transistors

QUICK REFERENCE DATA

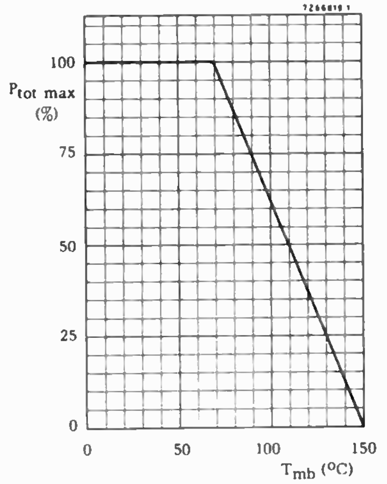
			BD135	BD137	BD139
Collector-base voltage (open emitter)	V_{CBO}	max.	45	60	100 V
Collector-emitter voltage (open base)	V_{CEO}	max.	45	60	80 V
Collector-emitter voltage (R_{BE} = 1 kΩ)	V_{CER}	max.	45	60	100 V
Collector-current (peak value)	I_{CM}	max.	1,5	1,5	1,5 A
Total power dissipation up to T_{mb} = 70 °C	P_{tot}	max.	8	8	8 W
Junction temperature	T_j	max.	150	150	150 °C
D.C. current gain		>	40	40	40
I_C = 150 mA; V_{CE} = 2 V	h_{FE}	<	250	160	160
Transition frequency at f = 35 MHz					
I_C = 50 mA; V_{CE} = 5 V	f_T	typ.		250 MHz	
Collector current (d.c.)	I_C	max.		1,0	
Total power dissipation to t_{mb} = 70°C	P_{tot}	max.		8 W	
Thermal Resistance					
From junction to ambient in free air	$R_{th\ j\text{-}a}$		100 °C/W		
From junction to mounting base	$R_{th\ j\text{-}mb}$		10 °C/W		
Saturation voltage					
I_C = 500 mA; I_B = 50 mA	$V_{CE\ sat}$			0,5 V	
D.C. current gain ratio of matched pairs					
BD135/BD136; BD137/BD138 BD139/BD140	h_{FE1}/h_{FE2}	typ.		1,3	
I_C = 150 mA; V_{CE} = 2 V		<		1,6	

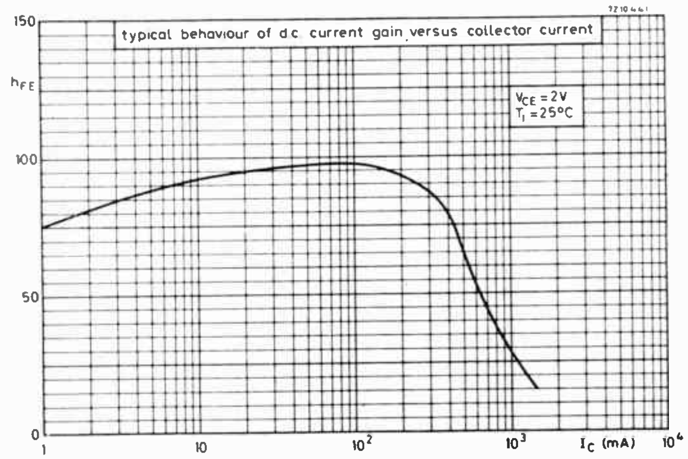
D.C. current gain			BD135	BD137	BD139
I_C = 5 mA; V_{CE} = 2 V	h_{FE}	>	25	25	25
I_C = 150 mA; V_{CE} = 2 V	h_{FE}	>	40	40	40
		<	250	160	160
I_C = 500 mA; V_{CE} = 2 V	h_{FE}	>	25	25	25

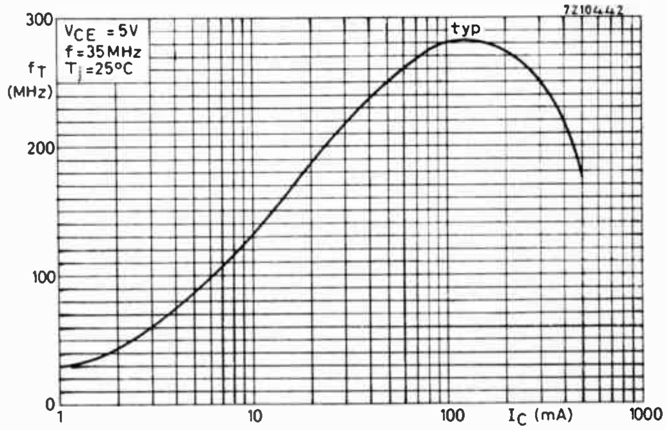
THE BD 135-139 SERIES ARE general purpose npn transistors recommended for driver stages in hi-fi amplifiers and TV receivers. They are housed in SOT-32 plastic cases. Their pnp complement is the BD 136-140.

Transistors BC635, 637 and 639 use the same chip as the BD135–139 series.

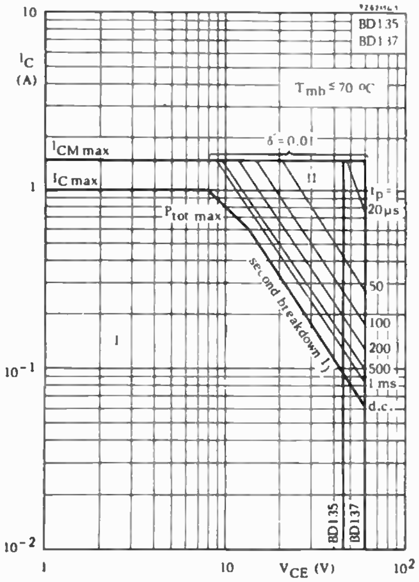
The devices are however mounted in a T092 plastic case. Power is limited to 1 watt — thermal resistance R_{th} j-a increases to 156°C/W.



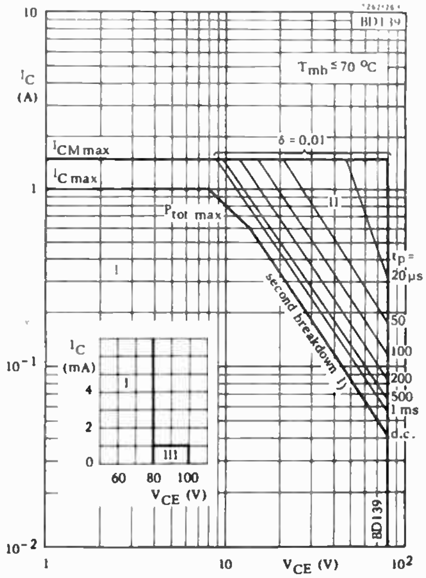




Maximum power dissipation related to case temperature.



Safe operating area for BD 135 & 137.

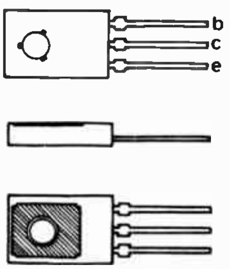


Safe operating area for BD 139.

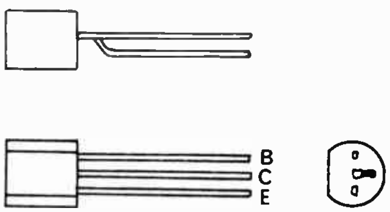
WARNING – operation outside the areas shown for longer than the time specified can lead to transistor failure.

Transistor Types				Connections
	BD135	BD137	BD139	1
Electrically similar (but see main text)	BD635	BD637	BD639	2

1



2



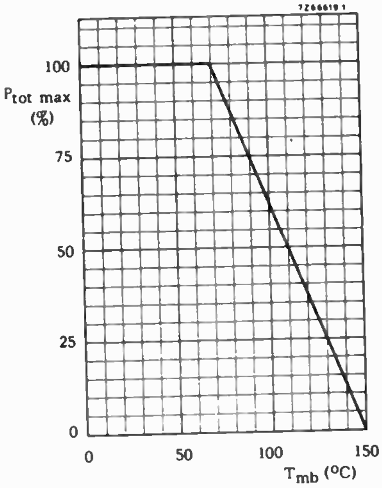
BD136-140 general purpose pnp transistors

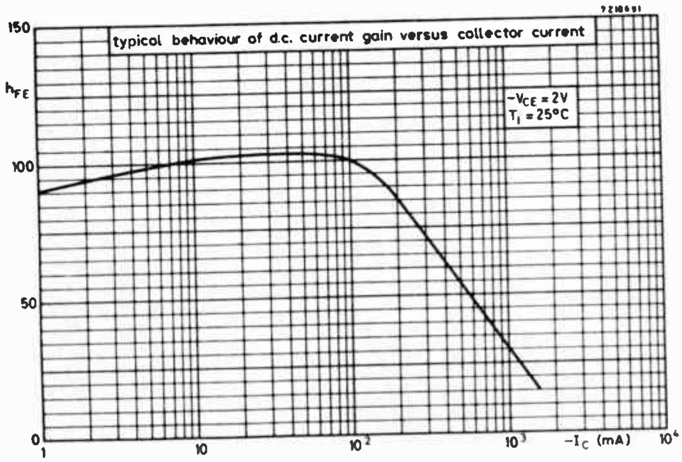
QUICK REFERENCE DATA

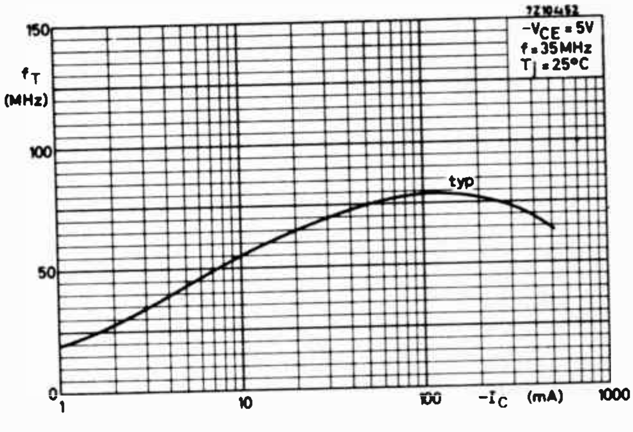
			BD136	BD138	BD140
Collector-base voltage (open emitter)	$-V_{CBO}$	max.	45	60	100 V
Collector-emitter voltage (open base)	$-V_{CEO}$	max.	45	60	80 V
Collector-emitter voltage (R_{BE} = 1 kΩ)	$-V_{CER}$	max.	45	60	100 V
Collector-current (peak value)	$-I_{CM}$	max.	1.5	1.5	1.5 A
Total power dissipation up to T_{mb} = 70°C	P_{tot}	max.	8	8	8 W
Junction temperature	T	max.	150	150	150°C
D.C. current gain		>	40	40	40
$-I_C$ = 150 mA; $-V_{CE}$ = 2 V	h_{FE}	<	250	160	160
Transition frequency at f = 35 MHz					
$-I_C$ = 50 mA; $-V_{CE}$ = 5 V	f_T	typ.		75 MHz	
Collector current (dc)	$-I_C$	max.		1.0 A	
Total power dissipation up to T_{mb} = 70°C	P_{tot}	max.		8 W	
Thermal Resistance					
From junction to ambient in free air	$R_{th\ j\text{-}a}$			100°C/W	
From junction to mounting base	$R_{th\ j\text{-}mb}$			10°C/W	
Saturation voltage					
I_C = 500 mA; I_B = 50 mA	$-V_{CEsat}$	<		0.5 V	
D.C. current gain ratio of matched pairs					
BD135/BD136; BD137/BD138	h_{FE1}/h_{FE2}	typ.		1.3	
BD139/BD140 $\lvert I_S \rvert$ = 150 mA; $\lvert V_{CE} \rvert$ = 2 V		<		1.6	
D.C. current gain					
$-I_C$ = 5 mA; $-V_{CE}$ = 2 V	h_{FE}	>	25	25	25
$-I_C$ = 150 mA; $-V_{CE}$ = 2 V	h_{FE}	>	40	40	40
		<	250	160	160
$-I_C$ = 500 mA; $-V_{CE}$ = 2 V	h_{FE}	>	25	25	25

The BD 136-140 series are general purpose pnp transistors recommended for driver stages in hi-fi amplifiers and TV receivers. They are housed in SOT-32 plastic cases. Their npn equivalent is the BD 135-139.

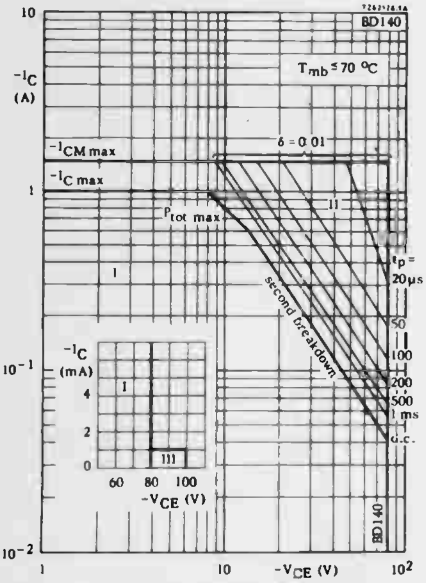
Transistors BD 636, 638 and 640 use the same clip as the BD 136-140 series. The devices are however mounted in a TO 92 plastic case. Power is limited to 1 watt and thermal resistance Rth j-a increases to 156° C/W.



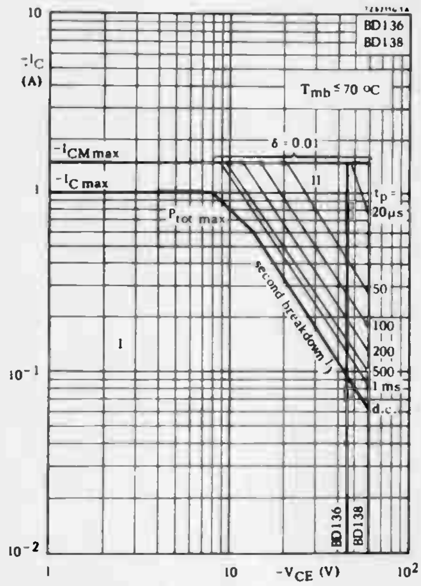




Maximum dissipation for related to case temperature.



Safe operating area for BD 136 and 138.

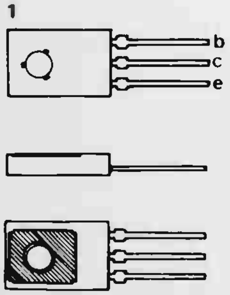


Safe operating for BD 140

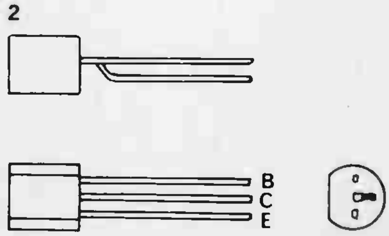
NOTE — operation outside the areas shown for longer than the specified time can lead to transistor failure.

Transistor Types				Connections
	BD136	BD138	BD140	1
Electrically similar (but see main text)	BD636	BD638	BD640	2

1



2



CIRCUIT DIAGRAM MARKINGS

ELECTRONICS Today International has adopted British Standard BS1852 : 1967 for marking component values on circuit diagrams.

The values of components are given by figures but the decimal point is replaced by a multiplier symbol in accordance with a table of standard prefixes. This procedure greatly reduces the possibility of errors (due to decimal points being left out, or a random printing spot falling in the wrong place).

Examples

4 k7	equals	4.7 k ohm
47 k	"	47 k ohm
1 m5	"	1.5 M ohm
4n7	"	4.7 nF
6p8	"	6.8 pF

Where a multiplier is not needed, the symbol 'R' is inserted to signify ohms.

Example

4R7 equals 4.7 ohms

Note also that capacitors that were formerly specified as decimal fractions of microfarads (10^{-6}F) expressed in nanofarads (10^{-9}F).

Example

0.01µF = 10 nF

Abbreviation	Read as:	Multiplies unit by:
T	tera	10^{12}
G	giga	10^9
M	mega	10^6
k	kilo	10^3
h	hecto	10^2
da	deka	10
d	deci	10^{-1}
c	centi	10^{-2}
m	milli	10^{-3}
µ	micro	10^{-6}
n	nano	10^{-9}
p	pico	10^{-12}
f	femto	10^{-15}
a	atto	10^{-18}

Standard prefixes . Multiplier symbols above 1000 are written with capital (upper case) letters, multipliers below 1000 do not use capitals (i.e. they are in lower case).

When spelled out in full, all multipliers start with a lower case letter (except when it is the first letter in a sentence).

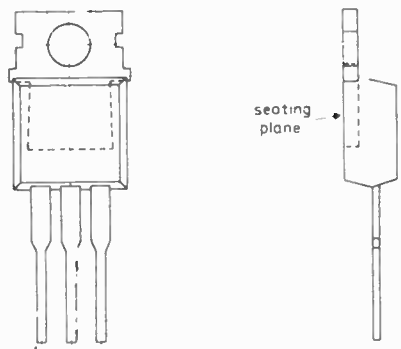
Thus — 10 MW = 10 megawatts
— 10 mW = 10 milliwatts

BD266-267 silicon power transistors

These are epitaxial base transistors in a monolithic Darlington circuit in a plastic TO220 package. This pair was used in the ETI 440 amplifier to provide high fidelity reproduction at 25 watts output.

The pnp types are BD266, BD266A, and BD266B; the npn types are BD267, BD267A, and BD267B.

TO220



BD266

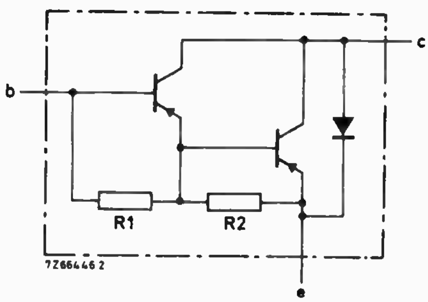
QUICK REFERENCE DATA

Ratings			BD266	BD266A	BD266B
Collector-base voltage (open emitter)	$-V_{CBO}$	max.	60 V	80 V	100 V
Collector-emitter voltage (open base)	$-V_{CEO}$	max.	60 V	80 V	100 V
Emitter-base voltage (open collector)	$-V_{EBO}$	max	5 V	5 V	5 V
Collector current (dc)	$-I_C$	max		8 A	
Collector current (peak value)	$-I_{CM}$	max.		12 A	
Base current (dc)	I_B	max.		150 mA	
Total power dissipation up to T_{mb} = 25 °C	P_{tot}	max.		60 W	
Junction temperature	T_j			150 °C	
Thermal Resistance					
(junction to mounting)	$R_{th\ j\text{-}mb}$			2.08 °C/W	
(junction to ambient)	$R_{th\ j\text{-}a}$			70 °C/W	
Characteristics					
Collector cut-off current (I_E = 0, V_{CB} = $-V_{CBO\,max}$)	$-I_{CBO}$	max.		0 2 mA	
T_j = 25 °C	$-I_{CBO}$	max.		2 mA	
Collector cut-off current (I_B = 0, V_{CE} = −30 V BD266) (I_B = 0, $-V_{CE}$ = 40 V BD266A) (I_B = 0, $-V_{CE}$ = 50 V BD266B)	$-I_{CEO}$	max.		0.5 mA	
Emitter cut-off current (I_C = 0, V_{EB} = 5 V)	$-I_{EBO}$	max		5 mA	
DC current gain (pulse conditions $t_p < 300\ \mu s$, $\delta < 2\%$) $-I_C$ = 0, 5 A, $-V_{CE}$ = 3 V	h_{FE}	typ		1500	
$-I_C$ = 3 A, $-V_{CE}$ = 3 V	h_{FE}	min		750	
$-I_C$ = 8 A; $-V_{CE}$ = 3 V	h_{FE}	typ.		750	
Base-emitter voltage ($-I_C$ = 3 A, $-V_{CE}$ = 3 V)	$-V_{BE}$	max.		2.5 V	
Collector-emitter saturation voltage ($-I_C$ = 3 A, $-I_B$ = 12 mA)	$-V_{CEsat}$	max.		2 V	
Transition frequency ($-I_C$ = 3 A, $-V_{CE}$ = 3 V)	f_T	typ		7 MHz	
C t-off frequency ($-I_C$ = 3 A, $-V_{CE}$ = 3 V)	f_{hfe}	typ.		60 kHz	
Switch-off second breakdown energy ($-I_B$ < 0)	$W_{(SB)}$	min		50 mJ	
Diode, forward voltage (I_F = 3 A)	V_F	typ		1 8 V	

BD267

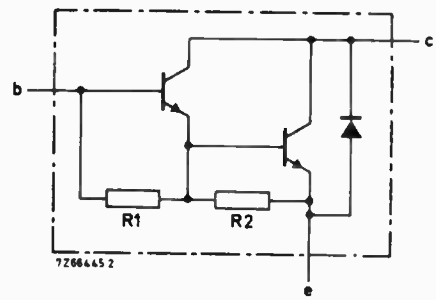
QUICK REFERENCE DATA

Ratings			BD267	BD267A	BD267B
Collector-base voltage (open emitter)	V_{CBO}	max.	80 V	100 V	120 V
Collector-emitter voltage (open base)	V_{CEO}	max.	60 V	80 V	100 V
Emitter-base voltage (open collector)	V_{EBO}	max.	5 V	5 V	5 V
Collector current (dc)	I_C	max.		8 A	
Collector current (peak value)	I_{CM}	max.		12 A	
Base current (dc)	I_B	max.		150 mA	
Total power dissipation up to T_{mb} = 25 °C	P_{tot}	max.		60 W	
Junction temperature	T_j	max		150 °C	
Thermal Resistance					
(junction to mounting)	$R_{th\ j\ mb}$			2.08 °C/W	
(junction to ambient)	$R_{th\ j\text{-}a}$			70 °C/W	
Characteristics					
Collector cut-off current (I_E = 0, V_{CBO} = 60 V for BD267, 80 V for BD267A, 100 V for BD267B)	I_{CBO}	max		0,2 mA	
T_j = 150 °C	I_{CBO}	max		2 mA	
Collector cut-off current I_B = 0, V_{CE} = 30 V BD267; I_B = 0, V_{CE} = 40 V BD267A; I_B = 0, V_{CE} = 50 V BD267B	I_{CEO}	max.		0.5 mA	
Emitter cut-off current I_C = 0, V_{EB} = 5 V	I_{EBO}	max.		5 mA	
DC current gain (pulse conditions $t_p < 300\ \mu s$, $\delta < 2\%$) I_C = 0, 5 A, V_{CE} = 3 V	h_{FE}	typ		1500	
I_C = 3 A; V_{CE} = 3 V	h_{FE}	min.		750	
I_C = 8 A; V_{CE} = 3 V	h_{FE}	typ.		750	
Base-emitter voltage (I_C = 3A, V_{CE} = 3 V)	V_{BE}	max.		2.5 V	
Collector-emitter saturation voltage (I_C = 3 A, V_{CE} = 12 mA)	V_{CEsat}	max.		2 V	
Transition frequency (I_C = 3 A, V_{CE} = 3 V)	f_T	typ.		7 MHz	
Cut-off frequency (I_C = 3 A, V_{CE} = 3 V)	f_{hfe}	typ		60 kHz	
Switch-off second breakdown energy (I_B < 0)	$W_{(SB)}$	min.		50 mJ	
Diode, forward voltage (I_F = 3 A)	V_F	typ		1.8 V	



R_1 typ. 6 kΩ
R_2 typ. 80 Ω

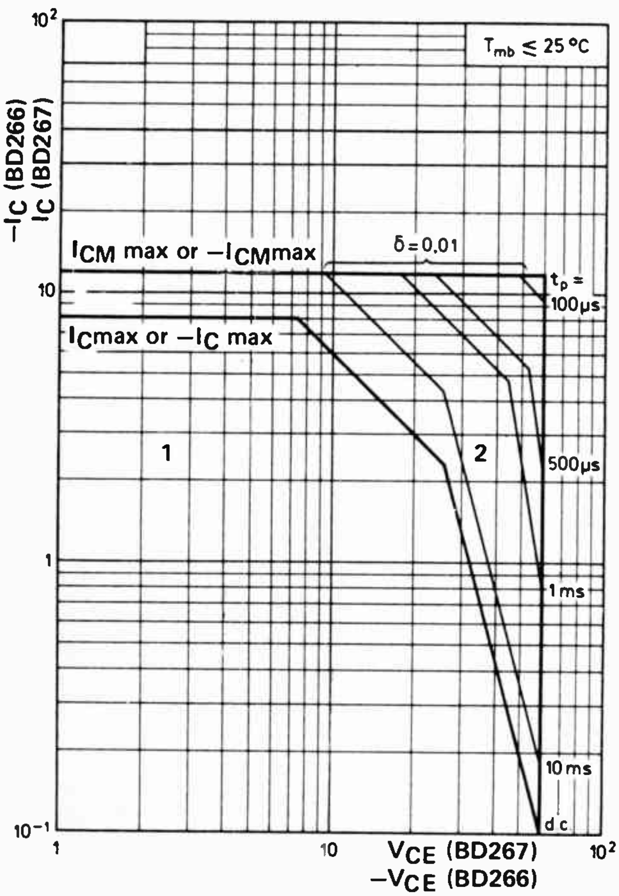
Internal circuit of the BD266
R1 is typically 10KΩ and R2 150Ω



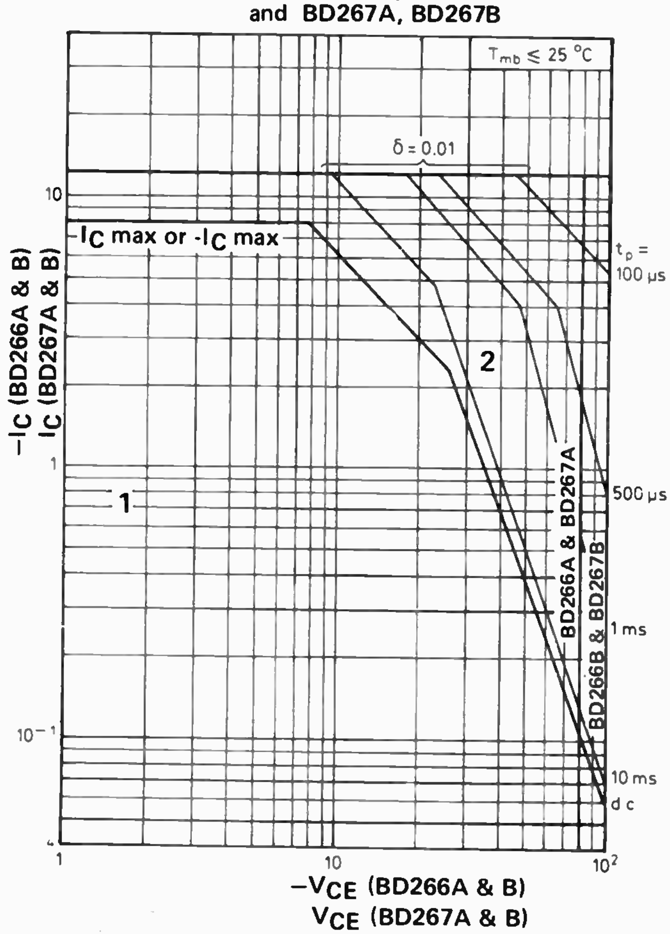
R_1 typ. 8 kΩ
R_2 typ. 100 Ω

Internal circuit of the BD267
R1 is typicall 10 KΩ and R2 150Ω

BD266 & BD267

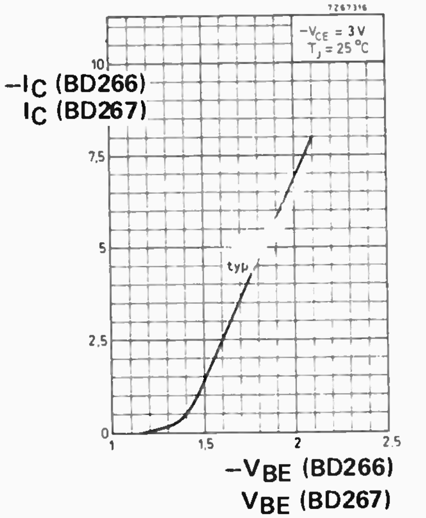


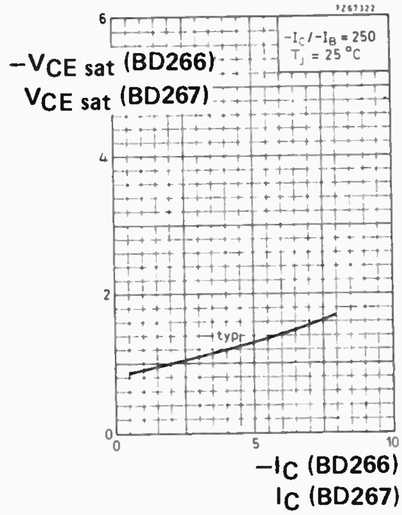
BD266A, BD266B
and BD267A, BD267B

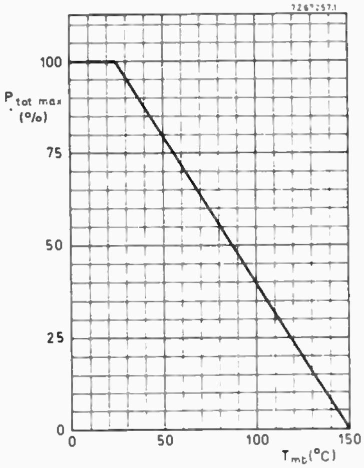


Safe operating area for the transistors forward biased. The area 1 is the safe region for dc operating. The area 2 is the region of permissible extension for repetitive pulse operation.

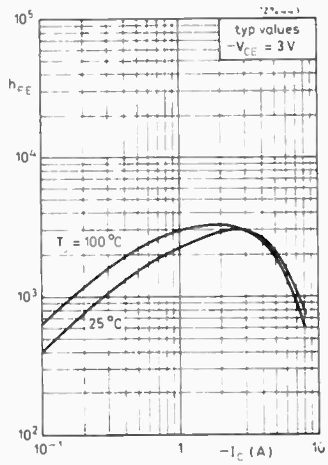
BD266-267 silicon power transistors







BD266, A, B



BD267, A, B