Low-speed differentiator

Monitoring slow changes in long-term experiments

by **L. Hayward,** Department of Geology and Mineralogy, University of Queensland

With certain electro-chemical experiments, it often becomes desirable to obtain the derivative of the output voltage/time curve in order that changes in the rate of change of amplitude become more easily observed. Such experiments often last minutes or even days, and consequently the classic type of RC differentiator seen in Fig. 1 is likely to be of little use, as the changes are so slow that great amplification is necessary, resulting in excessive noise masking the output.

This article describes an alternative form of differentiator, the block diagram of which is shown in Fig. 2. When read in conjunction with the timing diagram of Fig. 3, the operation is as follows.

A buffer presents the input signal to a pair of c.m.o.s. transmission gates. These are alternatively switched on for short periods, as determined by the clock generator and the sampling period monostable. The sampled voltages at t_1 and t_2 are stored in C_1, and C_2 respectively. The voltages across C_1 and C_2 are buffered by voltage followers, and applied to a differential amplifier. After t_1 and t_2, the resultant output from the differential amplifier is proportional to the difference of the charges on C_1 and C_2 that were set up during the interval t_1 to t_2. In other words $V_{(out)}=\Delta v/\Delta t$.

The timing diagram shows that, whilst the samples t_1 to t_2 and t_3 to t_4, etc., are of the same polarity, i.e. A-B, the periods t_2 to t_3 and t_4 to t_5, etc., give a reversal of polarity, i.e. B-A. Consequently, a further

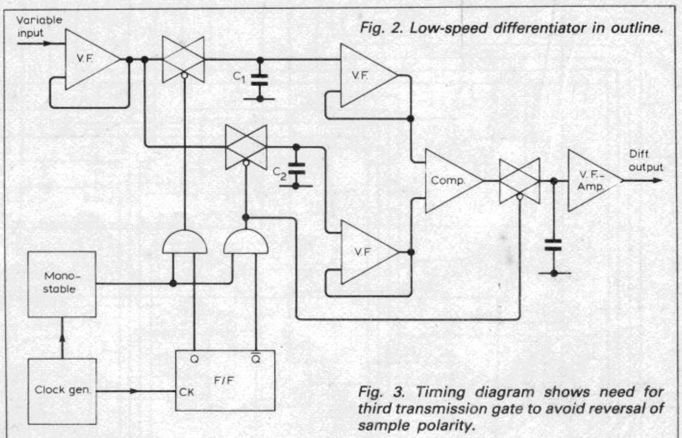
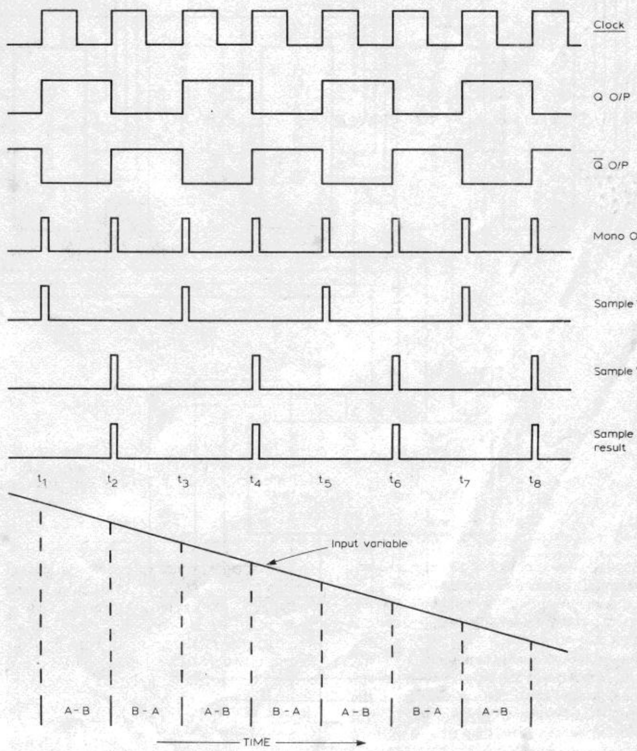


Fig. 2. Low-speed differentiator in outline.

Fig. 3. Timing diagram shows need for third transmission gate to avoid reversal of sample polarity.



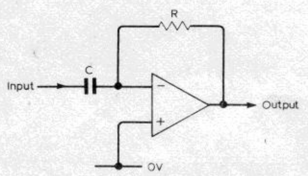


Fig. 1. Ordinary type of RC differentiator – useless for very long time intervals.

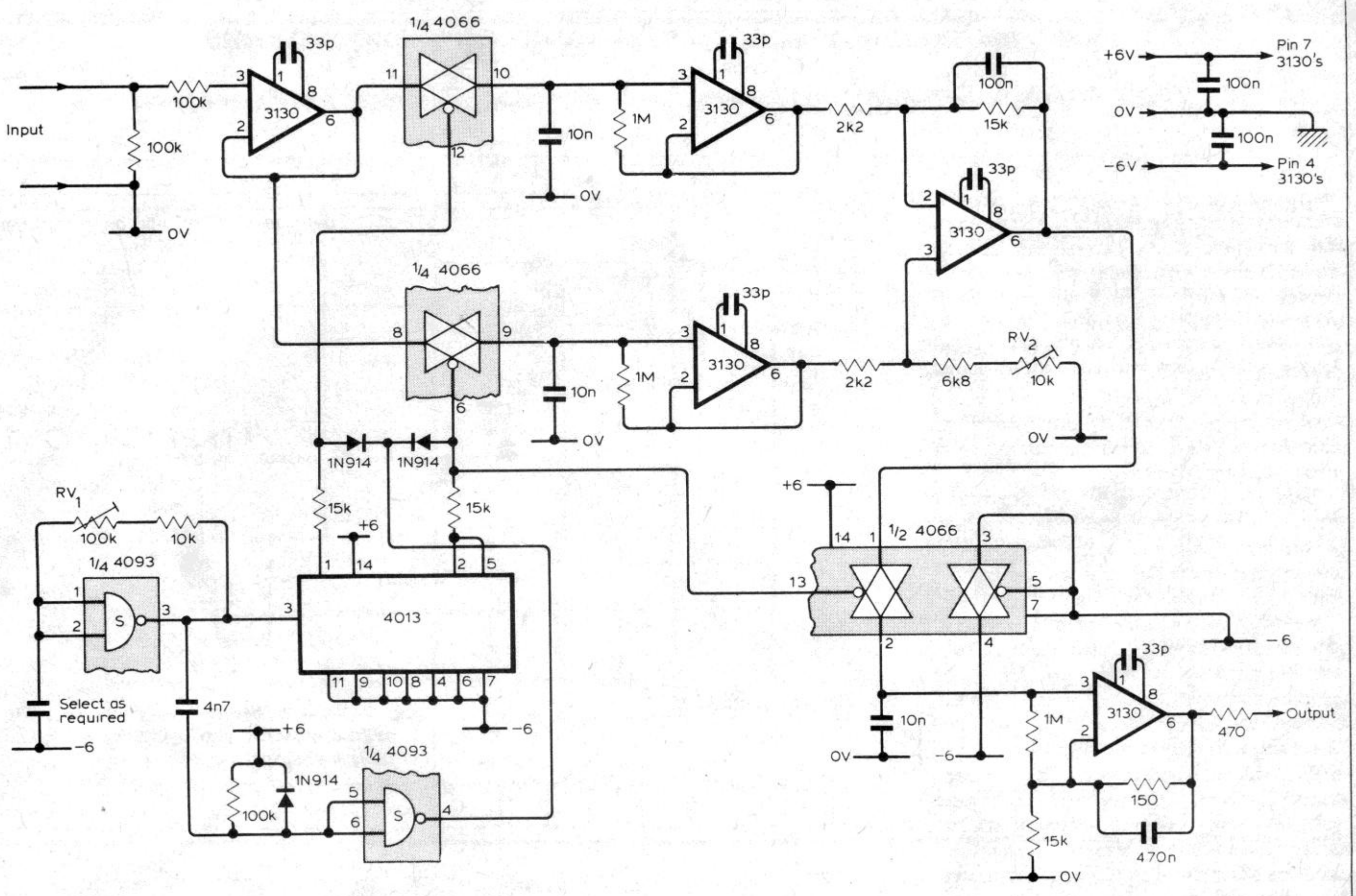


Fig. 4. Complete circuit diagram. 4013 is a dual, D-type flip-flop.

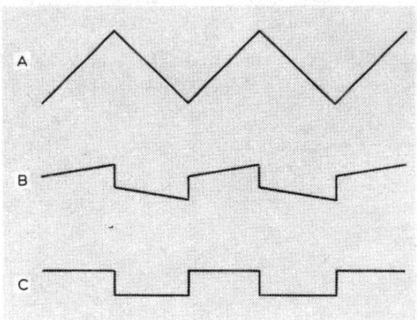


Fig. 5. The effect of adjusting RV_2 for differential balance. Triangular-wave input at (a) should produce square-wave output, as at (c).

sampling gate is required to eliminate the unwanted period. An output storage capacitor and output buffer complete the device, the complete circuit being shown in Fig. 4.

In operation, maximum sensitivity will be obtained when the clock frequency approaches the fastest rate of change of the signal. Clearly, the clock frequency should not be equal to, or less than this. The clock frequency is roughly adjusted by selection of capacitor, and fine tuned by the potentiometer RV_1. The only other adjustment is by RV_2 (differential balance). This is most easily set by observing the result of the triangle wave input (in Fig. 5). The output from the differentiator under these conditions should be a square wave, since we have a constant positive rate of change (gradient) followed by a negative gradient, and the amplitude of this square wave will be related to the input frequency. Set up RV_2 for maximum flatness of the square-wave output.

The circuit described is useful where a trend, rather than absolute results, are required. Clearly, this simple design could be elaborated to reduce offsets, and to use rather than eliminate the alternative sampling period, by more complex switching. Considering these limitations, the differentiator performs well and produces consistent results. □

Line scan camera

by L. Hayward
Technical director
Eastpoint Ltd

Considering the basic simplicity of linescan cameras – photosensitive array, lens and straightforward electronics – it is surprising they are not more widely used.

This article is based on results obtained using an EG&G Reticon LCO310 line scan camera together with some experimental electronics. The equipment was arranged to test the feasibility of a system for inspecting circular pressed metal lids passing along a conveyor. The report describes a practical system and suggests further work for improved operation.

The line scan camera operates like a normal optical television camera, with three important exceptions.

- Much higher stability, with no danger of burn due to continuous viewing of a stationary object.
- No field scan, so that instead of viewing an object using the raster principle, a single line only is viewed; consequently a movement of the item under inspection is required to view its complete area.
- The camera produces a 'clock' pulse for each picture element. This enables exact location of defects by counting clock pulses from start of scan.

The camera contains an array of photosensitive diodes (in our case 1024 but usually in multiples of 128 e.g. 256, 512 etc) mounted so that the image is focused onto it. Circuitry within the camera integrates the charge produced by the incident light, then clocks these representative charges out from the array by means of a shift register, thus providing a familiar video waveform. A sample and hold circuit provides a smooth transition from one element to the next. Electronics within the camera provide the clock timing pulses, and synchronized pulses for the analysing circuitry. The video waveform may be directly viewed on a oscilloscope to enable positioning, lighting, and focus.

As integration time varies with clock frequency, camera sensitivity to light varies. The amount of light required is greater the faster the camera is run. Maximum clock rate is 500,000 pulses per second, giving a minimum line scan time of 2.064ms, and 484 lines per second. From this it is possible to determine the maximum inspection rate, given required definition and object size.

Typical uses of such cameras are for determining overall dimensions of goods, for example paper width or box height. But the excellent stability and definition of the line scan camera should enable its use in much more detailed work.

Control electronics

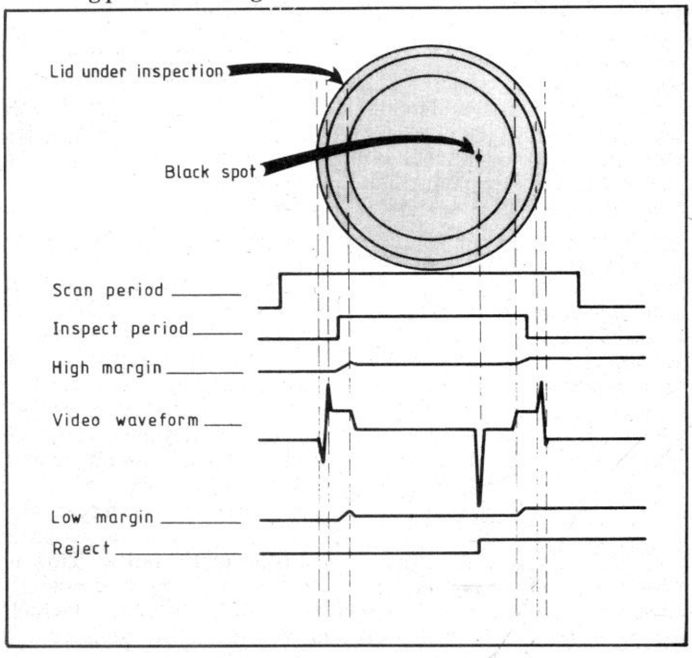
The high definition electrical representation of the surface under inspection requires suitable electronics to analyse the results and take decisions. The product being inspected is a range of lids having various sizes and coatings. The difference in size is fairly well accomodated by altering the field of view of the camera. It is clearly desirable that the maximum dimension of the product should always fill the viewing width of the camera to maximize definition. Experiments showed two main problems.

The extreme edge of the lid has a number of serrations, is folded at differing radii, and may be painted any colour. This makes total inspection of this area very difficult due to the randomness of the reflected light, and positional difficulty. This problem is still under investigation.

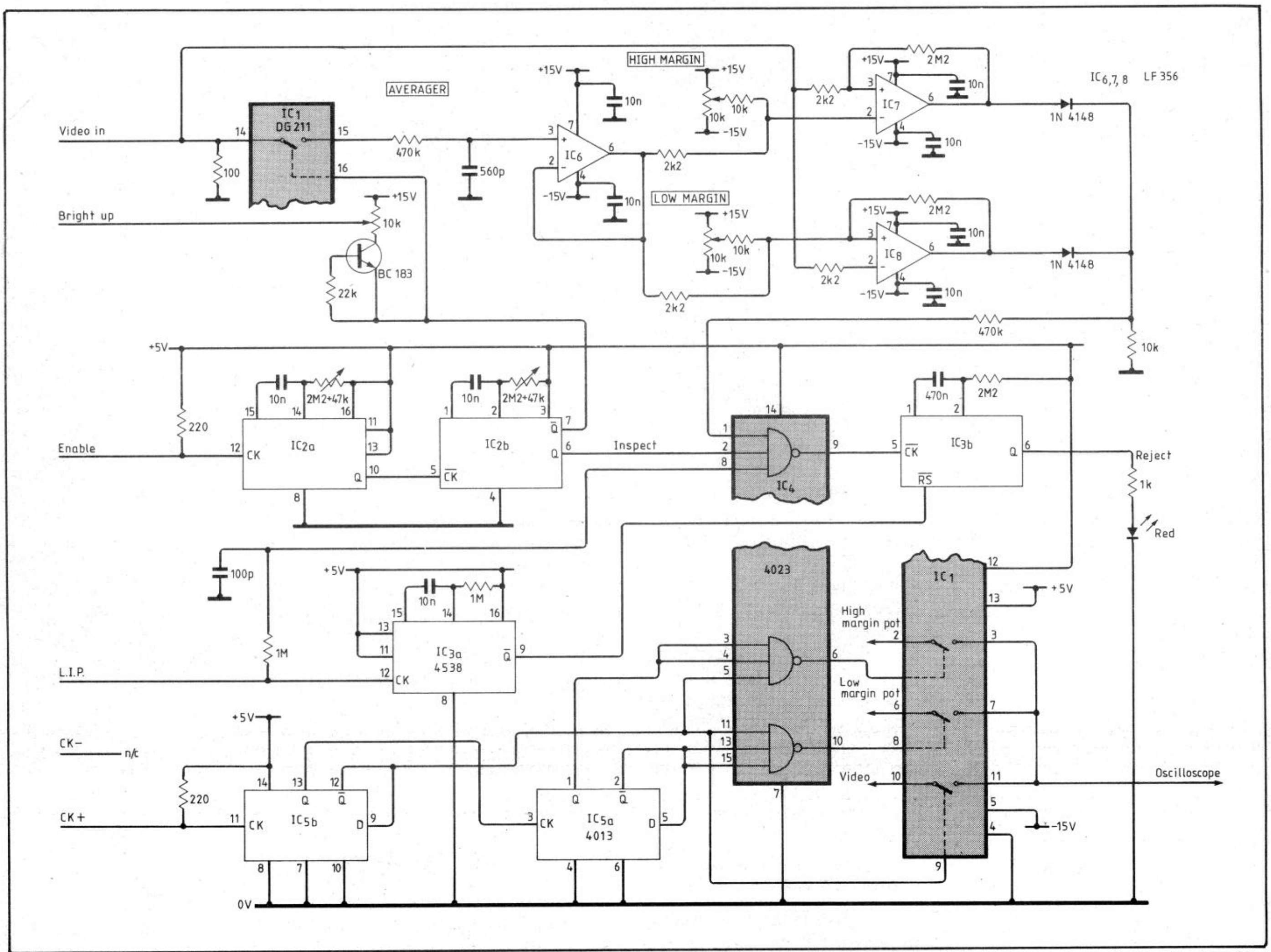
Lids may be coated with a light-coloured plastics material, which is easy to inspect, or not, in which case striations in the grain of the metal cause direct reflection of light from an unpredictable and wide range of angles. This should be overcome by using a very narrow angle of incident light with respect to the lid. At least two light sources will then be required to overcome shadow.

We constructed a form of optical bench that enabled lids of varying diameter to be placed at the camera's focal point. Using a local illumination of 50 watts at various angles and distances, the video waveform was observed. The definition was so good that the smallest visible blemish could be detected, and it is possible that 512 instead of 1024 pixels might be sufficient for the item in question, with a consequent saving of cost.

It was also clear from these tests that a fault might generate light levels either above of below that of the normal surface, particularly in the case of scratches, so that circuitry to detect this would be required. Circuitry to define a window of valid inspection would also be necessary to eliminate false rejection if the camera were scanning past the lid edge.



Black spot or scratch in this example of a bid returns strong reject signal.



Simple circuit detects flaws in material moving along a conveyor, provided that the inspected area is constant.

From these observations a basic detector unit was designed which works as follows. It is assumed that the lid can be fed intermitently, and rotated at least once in the camera field of view, and that a signal to indicate 'lid in position' can be derived.

The circuit of the detector electronics is shown above. Certain timing operations are generated by monostables. This was done for the sake of simplicity. A production unit would preferably have all timing derived from the camera clock for greater stability and ease of use. The camera is equipped to transmit differential data and clock signals, though this experimental circuit does not use differential receivers for the short range operation envisaged.

Operation

The video waveform from the camera is fed to two voltage comparators, IC_7 giving a positive output when the video is higher than threshold, and IC_8, when video is lower. The threshold voltage is an average of the video level, the waveform being integrated by the network at the input to IC_6. A voltage may also be injected from a high and low margin potentiometer to allow an additional offset in each case. Since the reference is in part derived from the average light level, reasonable variations in the light source are tolerable. The video waveform is only connected to the averaging circuit during the 'inspect' period, so that reflections from the edge and background do not distort the average.

At the start of each scan, the camera produces a positive going 'enable' pulse. This is used to trigger, in sequence, IC_{2a} and IC_{2b}, the first of which determines the non-inspect period from start to scan and the second determines the inspect period from then on. Two further monostables are used: IC_{3a} provides a delay to ensure a short scan period to allow the averager to settle after the lid is ready for inspection, while IC_{3b} is provided to light an indicator for about one second to indicate that a fault is present. The comparator outputs are OR-ed and gated with the 'lid in position' signal (LIP) The boolean expression for a reject then becomes

$$\text{REJECT} = (\text{COMP}_a + \text{COMP}_b).\ \text{LIP}.\ \text{INSPECT (after delay)}$$

It became obvious during early tests that a visual method of setting time and voltage margins would be required, and the remaining circuitry of Fig. 1 is designed to present all the necessary data on a single channel oscilloscope, having bright-up capability and sufficient speed (>10MHz bandwidth).

The camera clock drives a signal chopper, which splits the oscilloscope display up into a voltage displaying the high margin a voltage displaying the low margin and the video waveform. In addition the trace is brightened during the inspect period, so that its start and finish may be observed in relation to the video waveform. Page 49 shows a typical result.

A wideband buffer amplifier to drive the output to the oscilloscope is desirable for highest clock rate. Minimum times of the monostables will also need to be reduced. The trigger pulse for the oscilloscope time base is obtained directly from the camera, although the x-sweep could be generated directly from a digital-to-analogue converter driven by a counter of camera clock pulses.

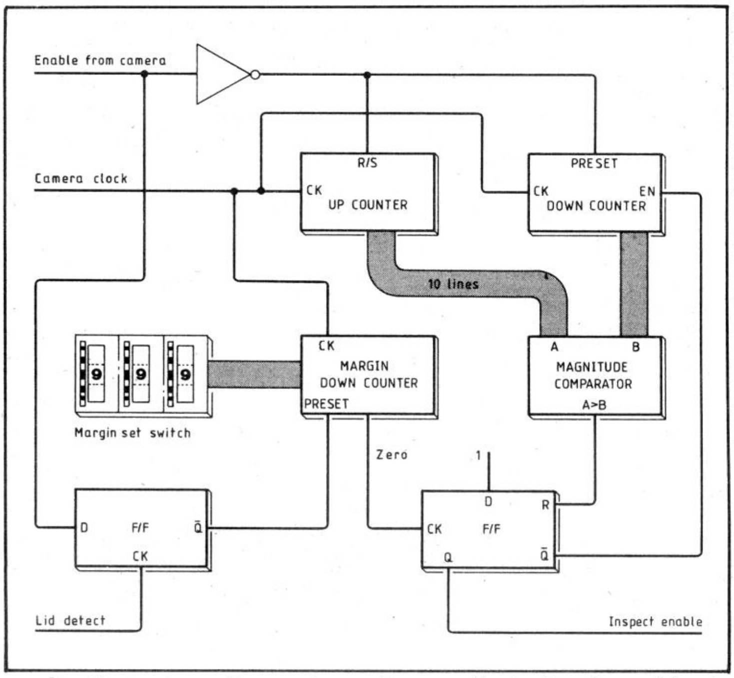
The result of many tests using this method has shown that a practical unit can be easily designed to fully inspect the interior face of lids of all practical diameters. The result of inspection using an intermittent and rotary scan would result in a slowing of the production line, but a linear scanning method would enable inspection at full conveyor line rate with minimal mechanical change. Problems arise regarding the initial sensing of the lid, tracking its edge, and then dynamically changing the non-inspect zones.

Sensing the lid edge as it appears in the camera field of view and adjusting the initial non-inspect area present little problem, but automatically defining the end of the inspect period is more complicated. The problem is not insuperable, and a block diagram of the proposed electronics is shown above. The theory assumes that the camera is adjusted in each case so that the field of view is that of the lid diametere, so that the lid diameter is 1024 elements. We can then state that the enable T_{on} measured in clock pulses from line start will be equal to the clock count a the detection of the lid edge plus the preset margin, and the enable T_{off}, will be equal to 1024 — the clock count at T_{on}. (The preset margin is assumed to be equal at start and finish of scan, and is thus included and defined at T_{on}.)

This hardware approach is preferred to a microprocessor application, since the system speed will always be limited only by the camera and not the computer processing time. In fact the hardware required should be about the same or less, and servicing easier.

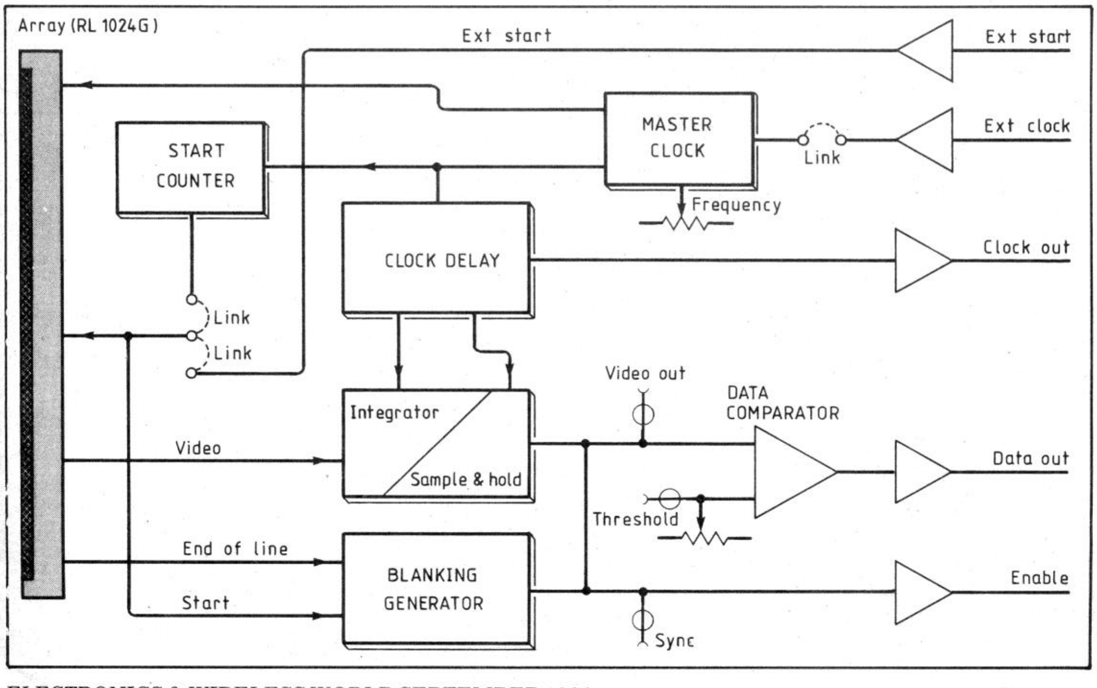
Clearly other techniques may be necessary to inspect other products. In cases where product speed is slow, inevitable with larger items, a microprocessor approach may be desirable to allow easier parameter change.

A variety of line scan cameras and arrays are advertised in this country and it may seem surprising that such a potentially useful device is not more widely used. One factor is the price. Diode arrays typically cost £500 for a 1024-element device, although some charge-coupled devices made by Fairchild would seem to give similar definition at a lower cost and faster data rate. ■



Inspecting circular objects where the inspect area varies according to diameter and position calls for a more complext digital solution.

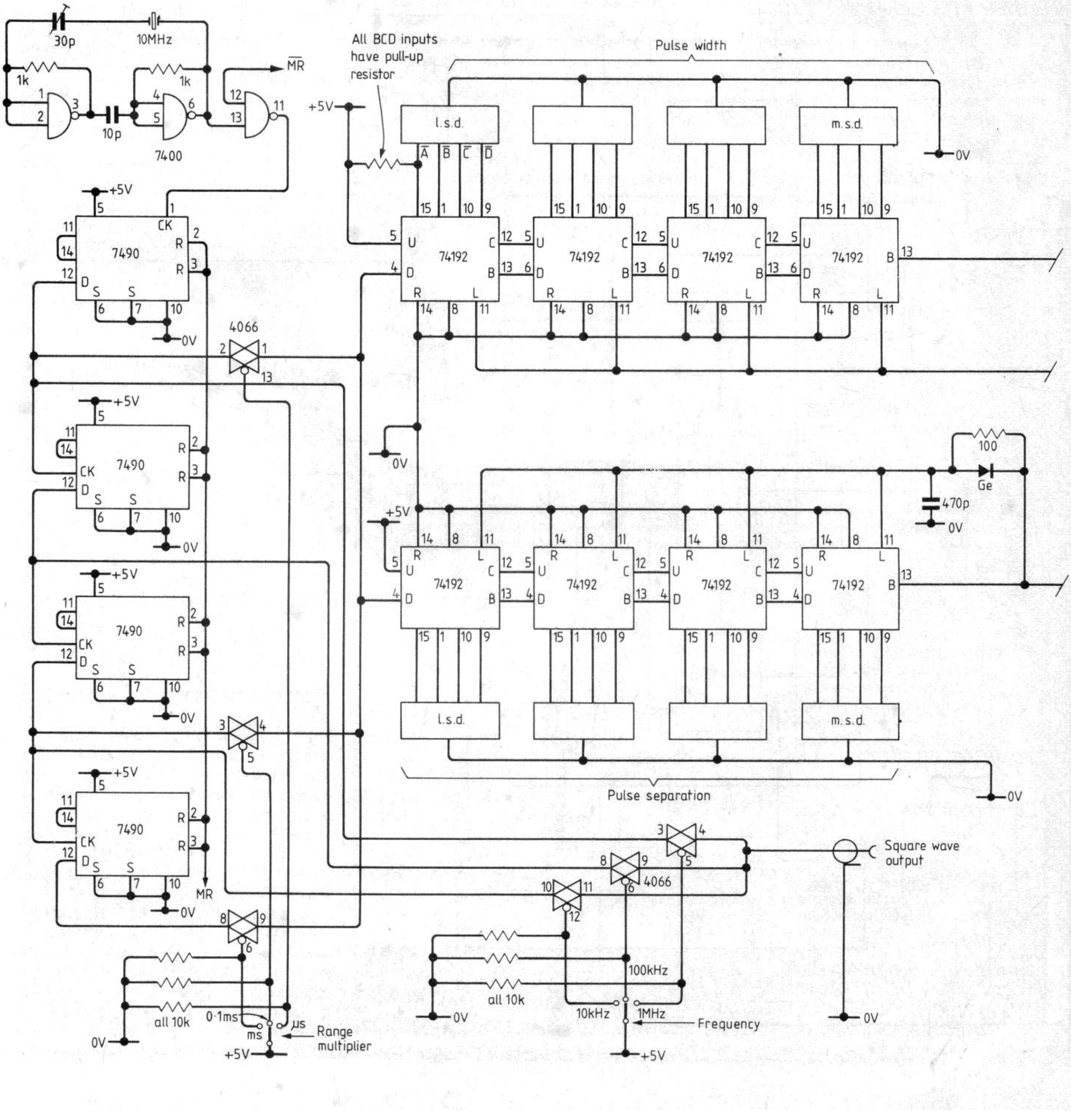
Only a simple circuit is required to produce video signal from the diode array. Power requirement is very low compared to vidicon cameras and overall stability much higher.



Precision pulse generator

by L. Hayward and G. E. G. Sargent

Standard pulse generators normally use an RC oscillator as a time-base and RC monostables to define the pulse width. This method gives a continuous tuning range and a simple circuit, but the accuracy is typically within 5% of the dial reading, and the pulse width and repetition frequency vary with temperature This design uses a crystal-controlled digital circuit to give high accuracy at a reasonable cost. With a suitable crystal, and oven, if necessary, the generator can be used to synthesize navigational or radar data, or to check computer systems.



This generator uses four thumb-wheel switches to select pulse width and separation from one microsecond to 9999 milliseconds in three ranges. The output is t.t.l. compatible and can provide any frequency in this range, i.e. $f = 1/t$. In addition, a standard-frequency square wave of 10kHz, 100kHz or 1MHz is available.

To prevent reflections along a coaxial connecting line, the source and receiver must match the impedance of the line. This is easily done at the receiver by padding the input with a suitable resistor. The transmitter should have its output also presented via a similar impedance, but the pulse must then be twice the required amplitude to allow for the voltage drop in a terminated receiver. Because this arrangement can lead to an accident if the termination is inadvertently disconnected and over-voltage occurs, a compromise is made. When the output is low, the line transmitter is terminated with 75Ω, and

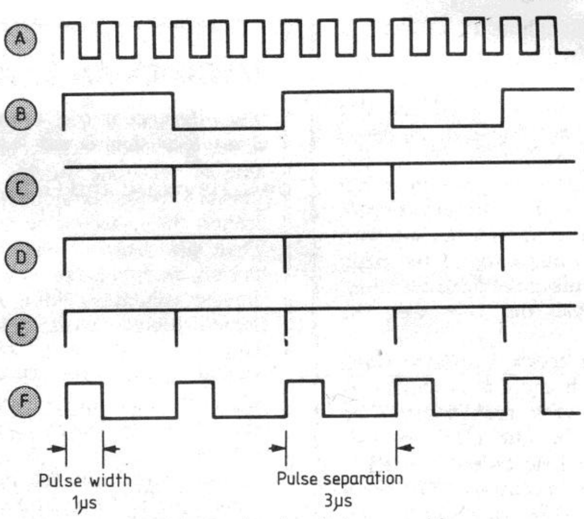


Fig. 1. Complete circuit. The thumb-wheel switches provide complement b.c.d. outputs.

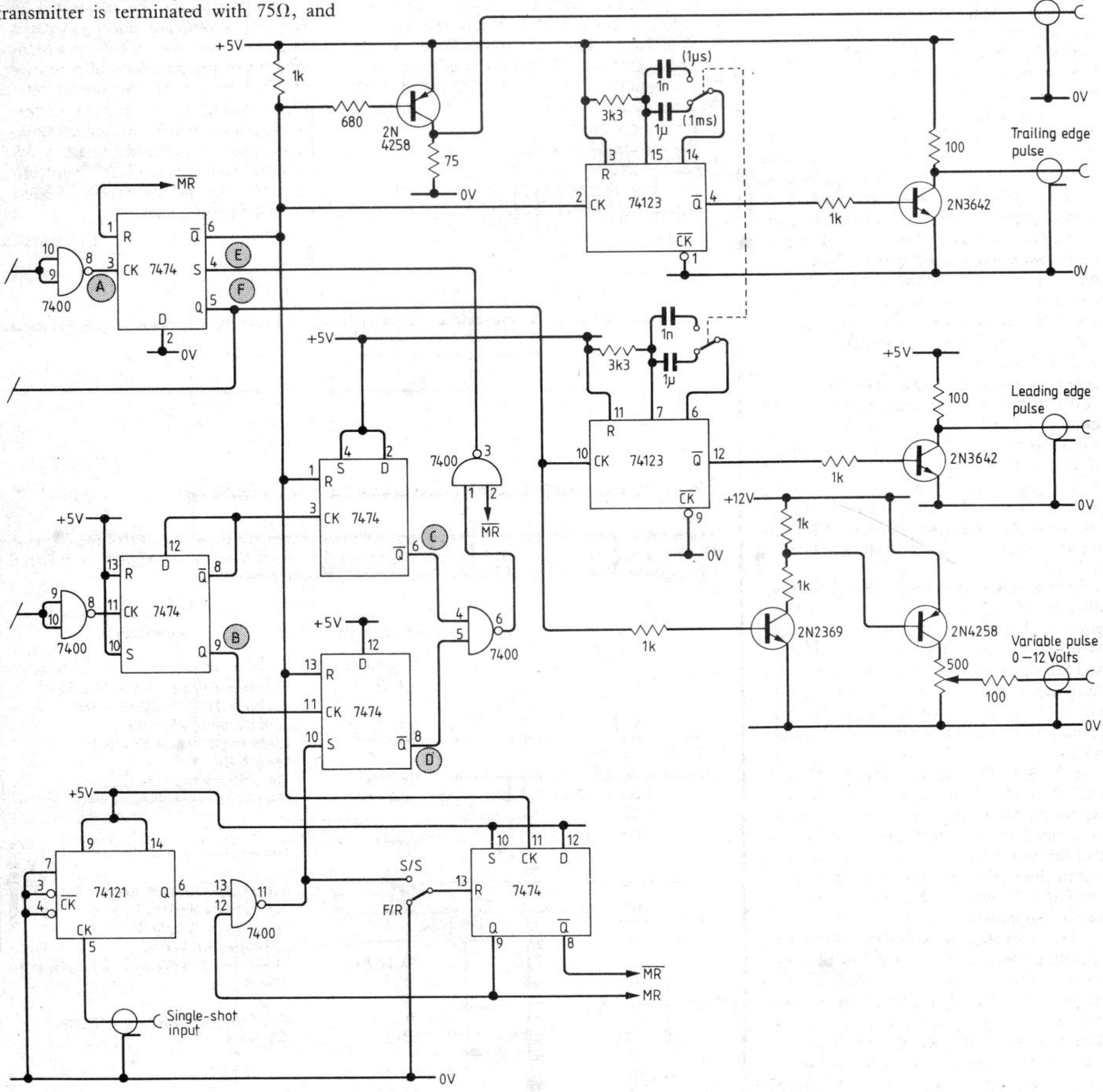


Fig. 2. Timing diagram.

DOCTORING POPS

L. HAYWARD* DESCRIBES TECHNIQUES FOR PRODUCING PHASING EFFECTS

THE term 'phasing' is used by recording engineers to describe an effect now being applied to some popular music. The sound is not easy to describe but resembles the fading heard on some distant radio transmissions. This effect can be brought about by introducing a notch in the frequency response of the recording equipment. If this notch is moved up and down the spectrum (**fig. 1**), the desired effect will be achieved.

The equipment required to accomplish this effect would be complex if based on electronic techniques. The four methods to be described are quite simple, requiring only conventional tape recording equipment, and form the basis of equipment employed in studios.

A really simple and effective method requires two recorders, each of which must be capable of simultaneous playback while recording. The inputs and outputs are combined and are in phase; if this is not so the outputs will cancel each other. With the machines running, one of the tapes should be slowed slightly by placing the thumb on the supply reel. If a signal is being presented to the inputs, portions of the frequency spectrum will be cancelled as the tape slows and regains speed.

The system in **fig. 2** is based on a single stereo recorder. (Some models have only a limited azimuth adjustment and may therefore not be quite satisfactory.) The input is fed to both channels and, while the tape is running, the azimuth of the recording head is moved from vertical to one side and back. If the recorder has off-tape monitor facilities, the effect can be heard while the azimuth is shifted. In the case of machines with an integrated record/playback head, the tape must be rewound first.

In both cases the outputs must be combined, preferably by parallel connection of the upper and lower track heads, so that as the head is adjusted the phase advance or retardation will cause cancellation of parts of the spectrum. Moving the record head slightly off azimuth will cause the high frequencies to be cancelled, while a large azimuth shift causes cancelling at the lower frequencies.

Fig. 3 is probably the best method of all, since sufficient longitudinal movement can be obtained to cover the full audio spectrum. Some engineering will be required, however, since normal tape transport systems are not designed to allow longitudinal adjustment of heads. The operation is the same as in **fig. 2** except that the head is moved from side to side instead of being rocked through azimuth.

On reflection, an easier method of obtaining the effect would be to alter the tape travel distance instead of the head spacing (**fig. 4**). The original idea of using two machines, for the record, was suggested to me by the BBC Radiophonics Workshop.

It should be noted that the tape speed in all cases should not be below 19 cm/s and that results are most satisfactory on recordings with a wide bandwidth.

**Haybridge Electronics*

FIG. 1

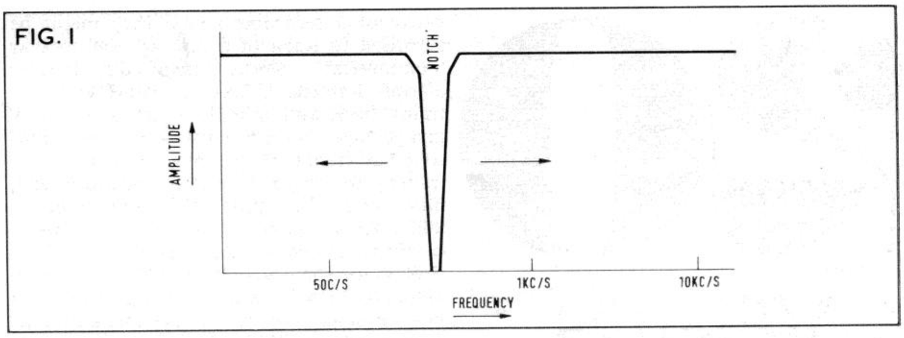


FIG. 2

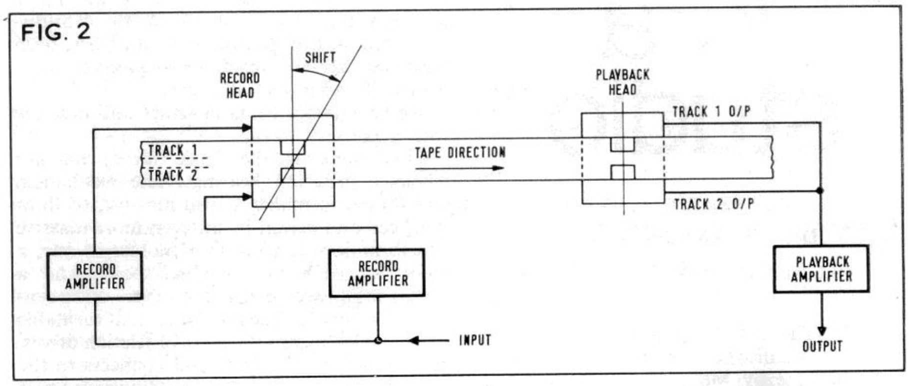


FIG. 3

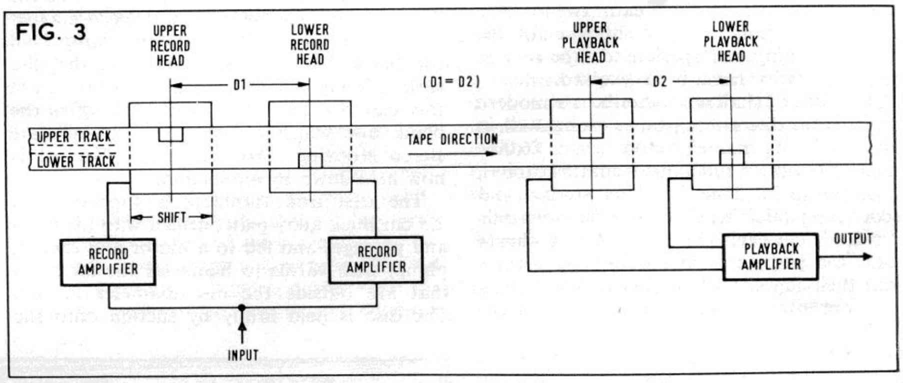
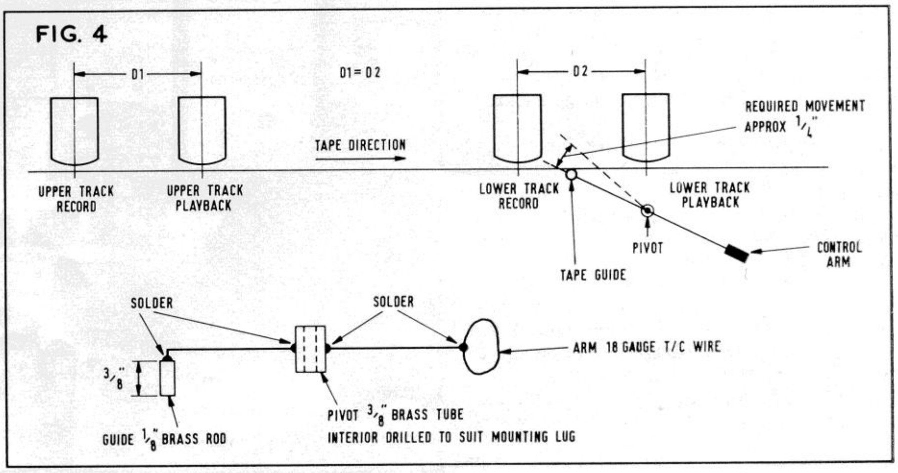


FIG. 4



A Versatile Recording Amplifier

Part Three **By L. Hayward**

THIS amplifier was developed to meet a requirement for a high quality playback system, particularly for use with the modern high output tapes. Such a system needs a good dynamic range and low background noise. These parameters are well met by the amplifier to be described.

The circuit is a derivation of the well-known 'feedback pair' (**fig. 1**) and this particular arrangement featured in a *Wireless World* article*. The use of the third transistor Tr3 (in **fig. 2**) allows the feedback circuit to remain isolated from the output, and the emitter-follower Tr4 reduces the output impedance sufficiently to provide a maximum output level of 0 dBm (1 mW into 600Ω). Silicon low-noise transistors are used. These are currently available types and have the added advantage of being cheap. As will be seen from the test figures, noise and dynamic range are quite satisfactory and there is a voltage overload margin of some 8 dB above the peak level produced by fully modulated EMI *815*. Due to the large degree of feedback, the distortion remains very low up to the clipping point, typically below .02 per cent at 1V rms output.

The capacitors C3 and C8 are added to improve hf stability and also to remove the possibility of rf signals being picked up and amplified by the circuit. The input coupling capacitor C2 is made much larger than is necessary for audio frequencies. This is to damp the input circuit of the amplifier when the head is connected, removing sub-audio noise.

The equalisation controls provide for adjustment at the ends and at the C/R turnover points in the characteristic curve. The values shown are for the CCIR 35 μS curve (38 cm/s). When correctly adjusted, it will be found that the combination of R18, VR19 and C11 does not exactly equal 35 μS. This is because the impedances associated with Tr1 and Tr3 are effective in the total time constant. The L/C adjustment L1, C10, was calculated to obtain an extended response from the head used for testing, a Bogen type *UK200*. It is resonant at a frequency of 20 kHz. For professional use this would normally be tuned to a lower frequency, and the amplitude control VR17 used to allow for gradual head wear. Since the range of adjustment for operation at lower frequencies, C11 is shunted by R20/VR21. This adjustment is effective in the region 35 to 50 Hz. If this were omitted a resistor across C11 would still be required to limit the gain at sub-audio frequency and preserve a good noise figure.

The adjustment of the equalisation controls is carried out as follows: Set the controls to a central position. Replay a test tape at 1 kHz. Adjust the output amplitude control to the level required, note this, and adjust VR19 for the same level at 10 kHz. Then VR21 at 40 Hz. L1/VR17 is set at the top end—the amount and frequency required will depend on the particular type of head used. A check should now be made of the complete range, adjusting slightly as required to obtain an optimally flat response.

The prototype was tested, and the results were as follows:

A range of frequencies was recorded on EMI *815* tape, using a Bogen *UK200* head. No recording pre-emphasis was used, and the head was fed from a constant-current source. The bias frequency was 100 kHz, set for 1 dB over-drop at 1 kHz. The tape was replayed from the same head to the amplifier on test. After setting up, the range from 35 Hz to 20 kHz was found to be within –1 and –2 dB, relative to 1 kHz. It should be emphasised that, since a compromise was made by using the same head for recording and replay, much greater accuracy can be expected when using a proper replay head and an accurate test tape.

FIG. I BASIC FEEDBACK PAIR

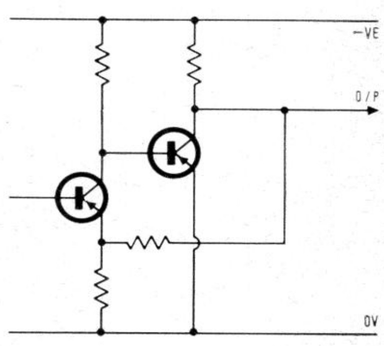


FIG. 2 PLAYBACK AMPLIFIER

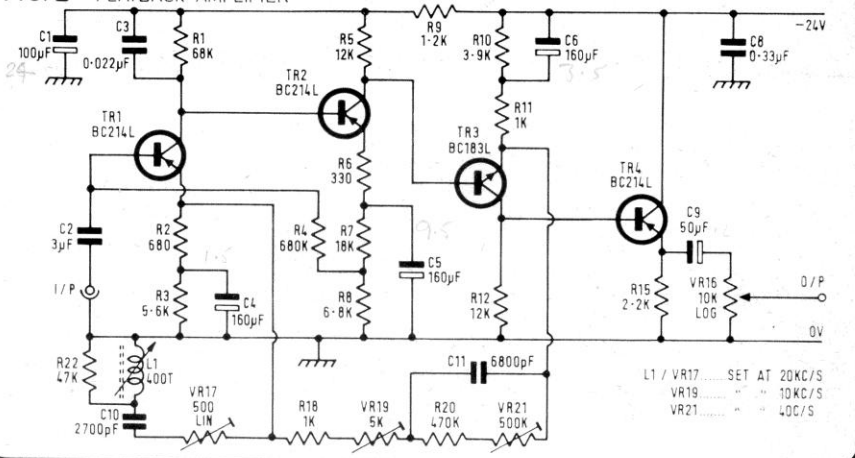
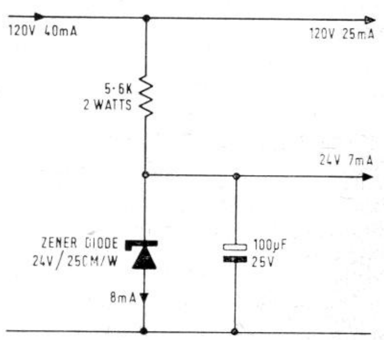


FIG. 3



The noise was measured with a mean sensing rms reading valve voltmeter having a response from 10 Hz to 10 MHz, and was found to be 72 dB below peak signal level with the input short circuited. With the head connected, this rose to –68 dB, mainly due to induced noise and hum in the long test lead and head.

Construction is not critical, provided that the amplifier is enclosed in a fully screened box, and is fed with a very well smoothed 20 to 27V supply. The most expensive component is L1, costing around £1. Many other types of pot core are suitable and, if heads other than the *UK200* are used, L1, C10 and R22 will require

(*continued on page 21*)

A Versatile Recording Amplifier

Part One By L. Hayward

THIS is the first article in a series of three. The remaining two circuits will be published in succeeding months and will consist of a bias and erase oscillator followed by a playback amplifier.

This design is intended to give a high standard of performance, particularly with high-output tapes such as EMI *815*. The main points of note in the design of such an amplifier are: that it should supply sufficient power to the record head to saturate the tape while itself contributing little distortion; and that it must deliver this power at a constant current, the reason for this being the variation of head impedance with frequency. If a constant current were not supplied, then the flux produced would be dependent upon frequency.

The simplest method of producing a constant current feed is employed here—namely, feeding the head via a resistor of sufficiently high value that the maximum current is determined by the resistor rather than the head. A few years ago, the disadvantage of this method was the shortage of transistors having a high enough V_{ceo} rating to provide the required voltage swing. No such problem now exists. Of course it is possible to produce a constant current by other methods, requiring less volts, but it must be remembered that, because of bias, upwards of 100V may be present across the head and this must be removed from the amplifier by some form of filter. The system finally adopted (emitter-follower with high resistor feed) is ideal for bias rejection, since the bias signal will be reduced by a factor of the series resistor over the amplifier output impedance, and is thus so small that it can be ignored.

Fig. 1 shows the basic circuit, clarifying the feedback paths. It was decided that the practical amplifier (**fig. 2**) must tolerate the swings of power supply voltage which are normally encountered when using direct mains-derived power, avoiding the need to stabilise the +120V ht rail. In order to do this, a large amount of dc and ac feedback have been incorporated. The overall signal gain has been set to saturate the tape from an input signal of 0 dB (about 800 mV). The bridging impedance of the input is a little less than 10 kΩ, the value of the level potentiometer. The maximum signal that the specified transformer will handle is +12 dB. The overall frequency response is set at 20 kHz by C6 and at 30 Hz by the transformer (–0.5 dB points).

No low level mic input stage or PPM has been included here, since these would normally be incorporated in a separate mixer, a subject very well covered in the recent series by David Robinson.

The amplifier is suitable for use with any tape or tape speed, and all but very low impedance record heads. The resistor R12 may be reduced to feed other than high impedance heads. The terminals marked xx in the circuit diagram refer to a suitable position for a series-resonant L/C equaliser if this is required. The values used will depend on tape speed, tape, head and bias, but an indication of the device is shown in **fig. 3**. This is similar to that used to make up for head losses in the playback amplifier which will be described next month, and resonates with the values described at 20 kHz.

The amplifier was tested using a Bogen type *UK200* head, EMI *815* tape, and bias set to 100 kHz and 1 dB overdrop.

It was found that about 30V peak to peak was required at Tr3 emitter to modulate the tape fully, and the amplifier output before clip point could be increased to 60V (+6 dB). This can be considered a quite good overload margin, since the distortion is sufficiently low until clipping occurs. An accurate distortion figure could not be produced, due to lack of instrumentation, but was checked approximately using the differential-null technique and found to be well below tape level. Noise was so low as to be ignored.

Care should be exercised in construction, in regard to keeping leads as short as possible. The transistors used have ft's well into the vhf region and will obligingly oscillate if allowed to.

It is advisable to mount the amplifier in a screened box to prevent interference pick-up. The only adjustment other than level which will need attention is the equaliser. This should be set to give an even hf response on playback, having first aligned the playback system accurately, with a standard test-tape.

FIG. 1 FEEDBACK PATHS

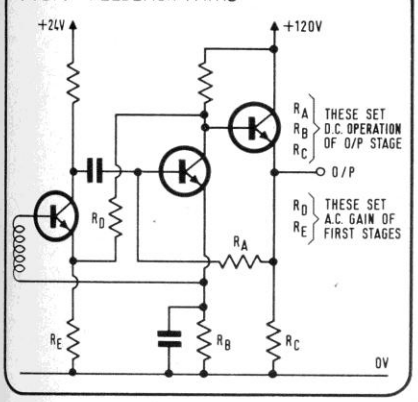


FIG. 2 CIRCUIT OF RECORDING AMPLIFIER

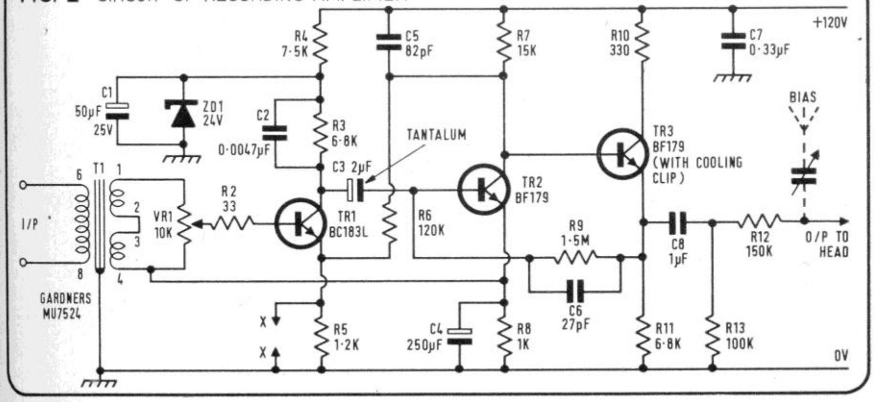
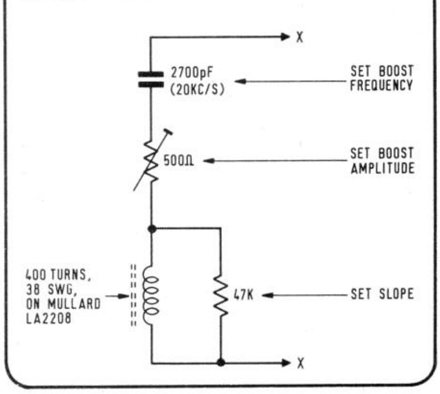


FIG. 3 EQUALISER



A Versatile Recording Amplifier

Part Two **By L. Hayward**

THIS article deals with what is often the weakest link, the bias and erase oscillator. No originality is claimed for this circuit; most engineers will be familiar with is, at least in valve form. Transistors lend themselves well to this type of multivibrator; this circuit has the advantage of simplicity and is easily designed to meet the power required for erasure. Although the system is to some extent self-stabilising, if half-track operation is required it is advisable to load the erase winding with the equivalent loss resistance of the unused part of the erase head. For the Bogen *UL290*, this works out at 10 k Ω. The best situation for good results is of course a full-track erase head, mono recordings being made using both half-tracks. If editing is to be carried out this is necessary in any case and leads to a much simpler construction.

The power developed in the heads will depend upon the power supply voltage, and this may require increasing if heads other than the *UL290* are used. To obtain sufficient erasure with a Leevers-Rich full-track erase head it was found necessary to increase the supply to 13.5V. It is unadvisable to increase this beyond 15V, however, since the peak collector voltage of the transistors may be exceeded. Where more or less volts are required for heads of an impedance other than the types specified, the secondary turns may be adjusted as required. As a guide, the *UL290* requires 70V rms. This is provided by 100 turns. It should be remembered that head manufacturers' specs regarding bias current often require updating, since the high output tapes in common use normally require a little more bias for 1 dB overdrop than earlier tapes. Where very low impedance heads are used it will be necessary to increase the gauge of wire used to accommodate the increased current. The operating frequency is set by Cx in conjunction with the total inductance, and in this case is set just over 100 kHz.

It is not advisable to run the oscillator unloaded, due to the high peak voltages produced. It is quite p●ssible to obtain an RF burn if fingers are placed across the secondary windings whilst operating in this mode!

The 500 μF capacitor is not included for decoupling but to ensure a gradual decay of oscillations when the oscillator is switched off. Its inclusion is therefore essential if magnetised heads are to be avoided. Bias amplitude control is effected by adjustable series capacitors feeding the heads, and the setting up of these will not shift the operating frequency to any great extent.

The pot-core assembly used here is, again, the Mullard *LA2208*. Thus the three circuits described in the series are standardised on this component. This was selected because its construction allows dismantling as often as required, making turns adjustment quite simple. The primary winding of the transformer should be bifilar wound. This is most easily accomplished by finding the length of wire required for one half of the primary, cutting two lengths, and then winding these on together. One start and one finish are now connected together to form the centre tap. The secondary windings may be pile-wound; one over the primary, the other in the spare half of the bobbin.

The complete oscillator should be built in a screened box to prevent radiation. Preferably the head feed resistors are also included in this box to improve bias rejection. These components were designated R12 in the record amplifier, and for the *UK202* head should be reduced to 82 kΩ. It is not advisable to use the higher impedance *UK200* with this circuit, as the required voltage leads to a rather large bias secondary. It can be accommodated, however, if the bias winding is connected in series with the erase winding.

It is advisable to use cooling clips of a minimum area of 25 cm^2 to mount the transistors, especially if greater output is required.

Either end of the power supply may be used as earth, since the transformer provides complete isolation.

Tests with the prototype have been most satisfactory. Using EMI *815* tape, adequate erasure was obtained with a variety of heads, including some 'difficult' types. Noise from the erased tape generally was no greater than 1-2 dB above virgin tape level, indicating a low distortion level from the oscillator.

Next month the concluding article in the series will feature a design for a high performance playback amplifier, as well as giving details of the power supply requirements for the set of three circuits.

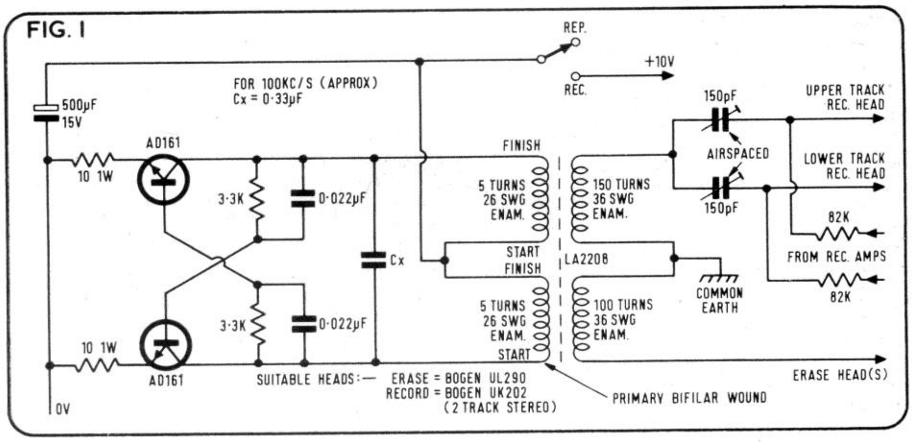


FIG. 1

Component specification

Z1	24V zener diode, 250 mW		
TR1	BC 183 L		
TR2	BF179		
TR3	BF179	With cooling clip	
VR1	10 k Ω potentiometer		
R2	33 Ω	10%	$\frac{1}{8}$W
R3	6.8 k Ω	5%	,,
R4	7.5 k Ω	3w	wire wound
R5	1.2 k Ω	5%	$\frac{1}{8}$W
R6	120 k Ω	,,	,,
R7	15 k Ω	,,	$\frac{1}{2}$W
R8	1 k Ω	,,	,,
R9	1.5 M Ω	,,	,,
R10	330	10%	$\frac{1}{8}$W
R11	6.8 k Ω	5%	2W
R12	150 kΩ or less depending on head		
R13	100 kΩ	10%	8W
C1	50 μF	25V Electrolytic	
C2	4,700 pF	polyester	
C3	2 μF	polyester or tantalum	
C4	250 μF	15V electrolytic	
C5	82 pF	5% silver mica	
C6	27 pF	,,	
C7	0.33 μF	polyester	
C8	1 μF	250V polyester	

Transformer. Gardners type *MU7524*, and international octal base.

Time sharing limited mains supplies

Automatic digital control for sharing mains power

by L. Hayward

If a number of loads on the same mains source requires more power than is available, some means of automatic power time-sharing is often the only practical way of avoiding blown fuses and overloaded generators and cables. The author was faced with the problem of using a 4kVA generator to drive a domestic cooker requiring 9kW with all its elements in operation. Although the design was originally intended for use in the kitchen it will be of value in any application where a number of cyclically heated devices are to be powered from a limited mains supply.

The problem that led to the designing of this circuit arose when a cooker rated at 9kW with all its elements in operation had to be driven from a 4kVA generator. Manual switching of the elements to share the power was possible but tedious and any mistake could have caused an overload, so some means of automatic power sharing was sought.

It was found that the elements of the cooker could easily be divided into three groups, each of which when driven alone would not overload the generator. All that remained was to design the following circuit to time-share the power between two or three of the groups automatically when the need arose.

Operating principle

The circuit is activated by switching on the cooker at its main terminal. To simplify the following explanation the term a 'group requiring power' is used to describe a group or part of a group which has been

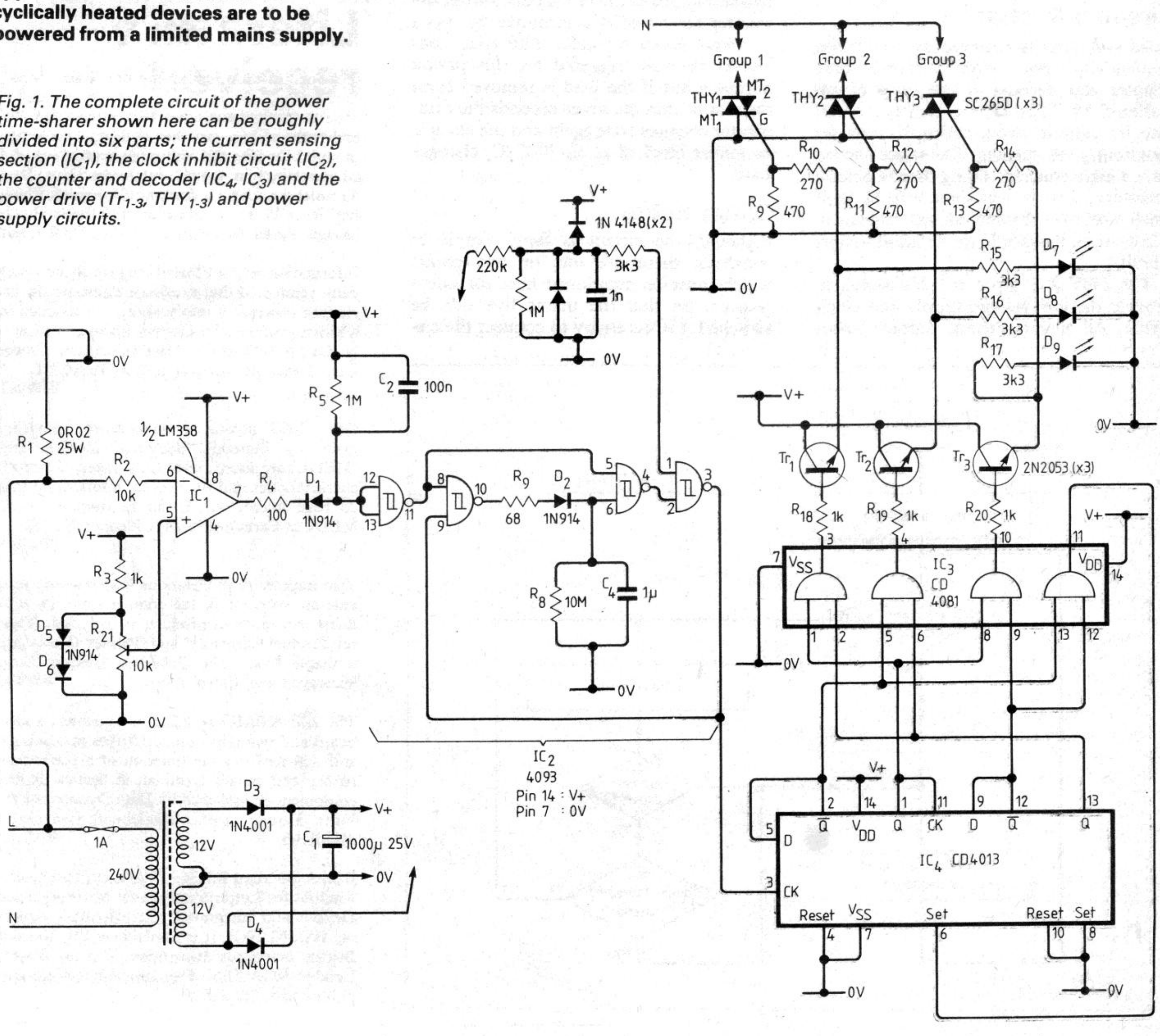


Fig. 1. The complete circuit of the power time-sharer shown here can be roughly divided into six parts; the current sensing section (IC_1), the clock inhibit circuit (IC_2), the counter and decoder (IC_4, IC_3) and the power drive (Tr_{1-3}, THY_{1-3}) and power supply circuits.

switched into the circuit either by the operator via the cooker control panel or by a thermostat.

When the circuit is activated but none of the groups require power each group is scanned in a 1-2-3-1 sequence to check whether it requires power. The scanning rate is determined by the mains frequency derived clock signal. When any group requires power it is detected by a current sensing circuit and the scanning stops. That group is then powered for around seven seconds and then the scanning continues. If the power requirement of the group ceases in less than seven seconds the scanning continues from that point. Because the scan time is very small in relation to each seven second power-up time full use of the available power is made however many groups are switched in.

To summarize, if only one group requires power it receives power constantly apart from the small scan time at the end of each seven second period. If two groups require power they each receive it alternately for seven second periods. When all three groups require power each receives it in turn for seven second periods according to the 1-2-3-1 sequence.

Circuit operation

Load switching is achieved by 40A triacs sequentially driven from a ring-of-three counter and decoder whose outputs are buffered by Tr_{1-3}. These transistors can also be used to drive optional l.e.ds for monitoring the outputs. The triacs chosen have a higher current rating than is strictly necessary, so that heat sinks may be kept small, and to withstand enough current to trip a circuit breaker in the event of a short circuit.

The 240V a.c. input is transformed to provide the low voltage supply and clock signal. All mains current output passes through R_1 which is made up of five 0.1Ω,5W resistors connected in parallel This low resistance value was chosen to reduce dissipation and avoid wasting power. When the voltage across this resistor exceeds a preset level the comparator output changes to logic 0 on every positive half cycle. The time constant of R_1, C_2 is chosen to integrate these pulses sufficiently to provide a constant logic 1 at the output of the first Schmitt trigger. When no current is drawn from the mains the first Schmitt trigger output will be low and the output of the third trigger (pin 4, IC_2) will be high and mains frequency clock pulses will be admitted to the counter, IC_4. This counter is a modified divide by four circuit.

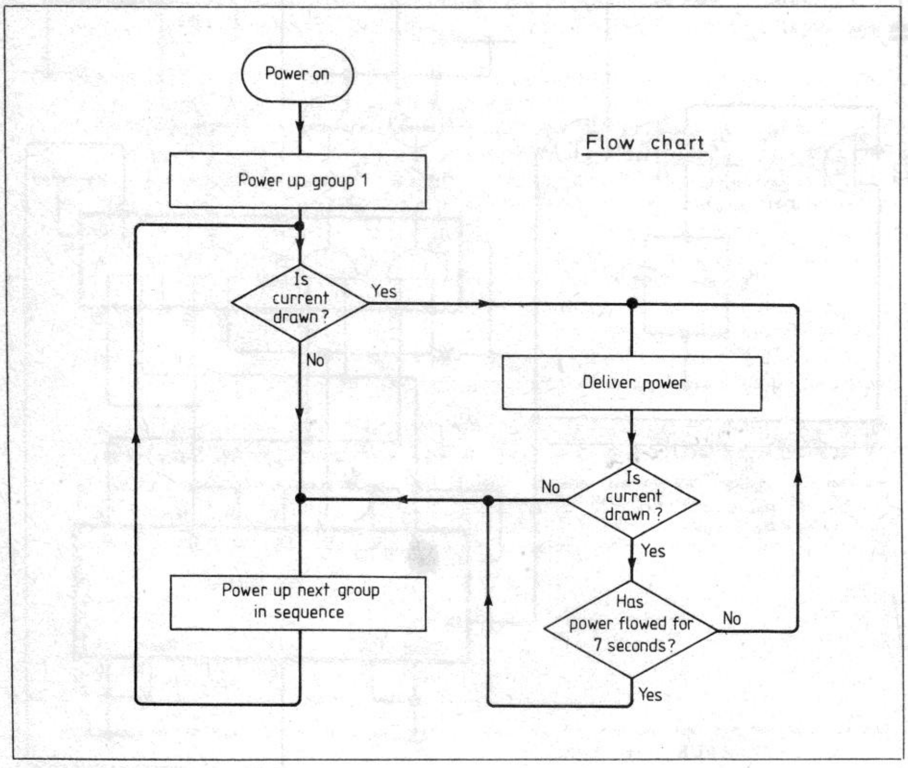
The decoder consists of four AND gates, the outputs of which go sequentially high as long as clock pulses are received by the counter. Three of the gates, when high, each trigger one triac through a buffer transistor. The fourth gate sets the counter to restart the sequence. If a load is connected to any of the triacs, the momentary triggering caused by the scan causes a voltage drop over R_1. The comparator changes state and the Schmitt trigger configuration blocks the clock pulses from the counter for a period determined by R_8/C_4 – about seven seconds. The triac concerned remains triggered for this period normally but if the load is removed from the triac within the seven seconds the comparator changes state again and the clock is no longer blocked as pin 4 of IC_2 changes state.

Construction

Although the circuit is fairly simple to construct there are one or two points which must be mentioned here for safety reasons. So that the mains live side be switched it is necessary to connect the circuit's 0V rail through R_1 to the live side of the mains. Consequently the whole circuit is live and must be enclosed to eliminate any possibility of accidental touching. The circuit must be completely isolated from its enclosure and from the chassis of the cooker. Don't forget that any heatsinks connected to the triacs will be live also. On no account should the circuit be wired such that current for the cooker is drawn through the printed circuit or Veroboard rails.

The only setting up required is the adjustment of VR_1 so that the minimum load is just sufficient to stop the counter.

In theory, if all three groups of the cooker are demanding power at the same time then the heating time per group will be three times longer than when only one group is used. In practice this has not been found to be a problem because cookers are rarely used to their ultimate temperature capability and any thermostatically controlled elements only switch on for brief periods once the set temperature is reached. □



Literature received

Microcomputer **analogue I/O** systems for several types of bus, together with data converter modules for lab. and industrial application, are all described in a new catalogue from Data Translation, Ltd, an American company which has its UK sales office at 430 Bath Road, Slough, Berks. SL1 6BB. WW401

Information on the Fiberfil range of flame-retardant, reinforced **thermoplastic compounds,** including processing information, is contained in a leaflet produced by Fiberfil Europe, c/o Capital Controls Division, Dart Industries, Crown Quay Lane, Sittingbourne, Kent ME10 3JE. WW402

Over 1600 **power semiconductor devices,** made by Marconi Electronic Devices Ltd (MEDL) are listed, with their salient electrical characteristics and mechanical information in a 26 page guide, which can be obtained from MEDL at Carholme Road, Lincoln LN1 1SG. WW403

Two impressive publications from **Plessey** provide an overview of the company and its products. An index of products is included. 'Plessey Product Directory' and 'Plessey Group' are available from The Publicity Liaison Unit, Vicarage Lane, Ilford, Essex. WW404

The ZIP ASR/K7 **terminal** is a printer, a keyboard and two mini-cassette drives in one unit, and will perform the function of a paper tape reader and punch terminal. A leaflet on the equipment is published by Data Dynamics Ltd, Data House, Springfield road, Hayes, Middlesex. WW405

British Standard BS3363:1980 is entitled 'Specifications for **Letter Symbols** for Semiconductor Devices and Integrated Microcircuits', replacing BS3363:1968. It is available at £16.50 from British Standards Institution, 2 Park Street, London W1A 2BS. Two supplements are also published at £12 and £9.

THE 'VIRTUAL EARTH' AS A MIXER

BY L. HAYWARD

FIG. 1

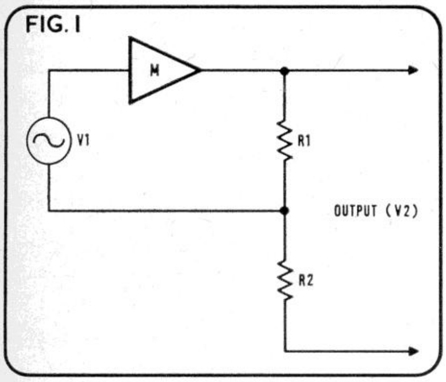


FIG. 2

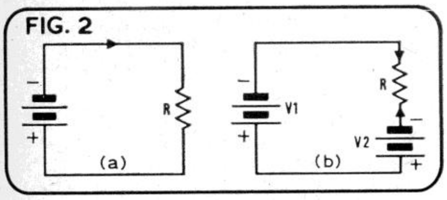


FIG. 3

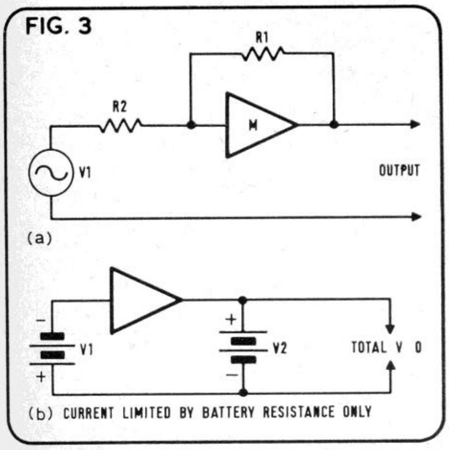


FIG. 4

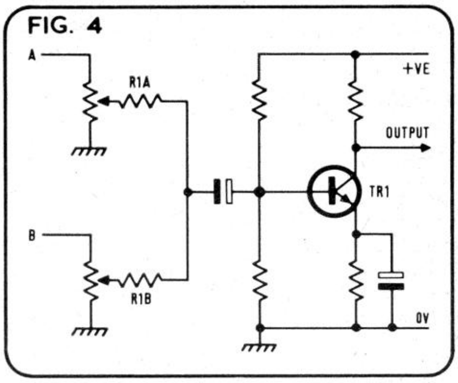


FIG. 5

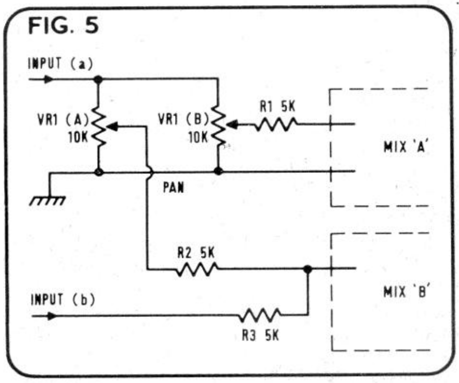


FIG. 6

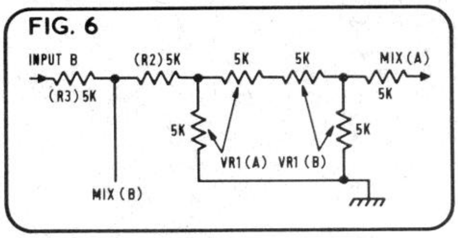
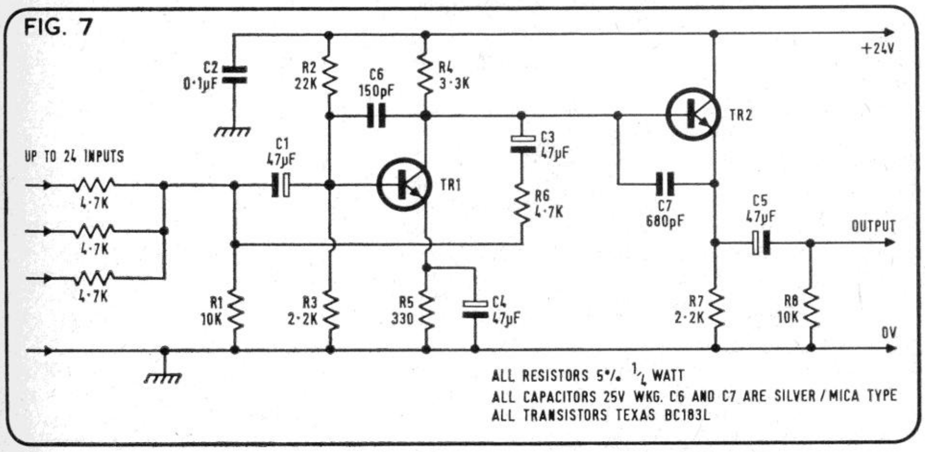


FIG. 7



IN order to understand the operation of the virtual earth amplifier, it is necessary to be familiar with certain basic facts of negative feedback. A brief and non-mathematical approach to the methods used will first be given.

When negative feedback is applied from output to input of an amplifier, the result will be an impedance change and a reduction in gain. **Fig. 1** shows an amplifier (M) where negative feedback is in series with the input signal. The open loop gain of M is assumed to be 100 times or greater. With feedback applied, the gain will approximately equal R1/R2 so that if R1 = R2 the gain with feedback will be unity. Since the feedback is in series with the input, the input impedance will be raised. A rough analogy of this is illustrated by **fig. 2.** This shows **(a)** a battery V1 driving a current through resistor R. Another battery of equal voltage, V2, represents a negative feedback voltage **(b)**. Little or no current flows in the second case, even though R has the same value, so the impedance 'seen' by V1 is high.

Fig. 3a shows an identical amplifier with (R1 = R2) unity gain. The feedback is now in parallel with the input signal, in effect shorting it out in the **fig. 3b** analogy. The total V1 + V2 is zero, due to their reversed polarity, but in this case the impedance is low since maximum current is flowing. It will be seen how the term 'Virtual Earth' is derived since the voltage at the junction of R1 and R2 will be approximately V1 minus M volts, and the input impedance very low. This point is, therefore, virtually at 'earth'.

The advantage of this circuit when used as a mixer can now be shown. **Fig. 4** shows an arrangement commonly used for mixing. The resistors R1a and R1b are chosen to prevent interaction of pot A against the signal level of B and vice versa. This has the advantage that a loss occurs in the mixing network which must be made up by the following amplifier with consequent extra noise. A well-designed system on this basis is tolerable, if only mono signals are being mixed, but what of the case with stereo?

Fig. 5 shows part of a stereo mixer. VR1 is a pan pot for mixing signal (a) to output A or B. Input (b) is fed only to the output B. Should the pan pot be left in a central position, a coupling exists between A and B, which is analysed in **fig. 6.**

Fig. 7 shows a suitable design for a virtual earth mixer. The impedance at the base of Tr1 is about 50 ohms due to negative feedback, via R6 and C3, and the rejection using this system can be as much as 70 dB between two similar channels. A further advantage is complete freedom of interaction between faders. Up to 24 channels can be mixed into this one circuit. The gain due to feedback is unity and the noise level better than –70 dB, with an input of 0 dB (allowing about 12 dB overload margin). Tr2 is added to provide a low output source impedance, although the overload margin will be reduced if this is bridged with 600 ohms or less.

The mixer is flat over the entire audio range and cut above 60 kHz by the inclusion of C6.

As a footnote to this article, readers' attention is drawn, when reading the specification supplied with commercial mixers, to the figure given for stereo separation. In most instances, it is not even mentioned!

NRZ RECORDING FOR SMALL COMPUTERS

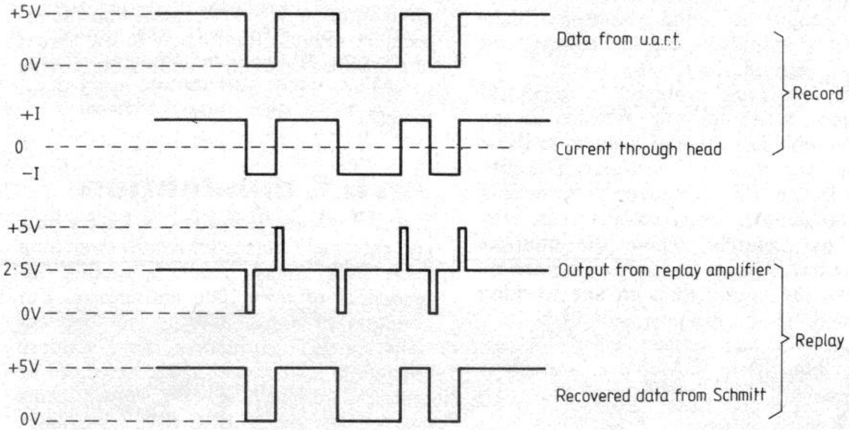
The majority of small computer recording systems use the Kansas City cassette recording format, with a data rate limited to a few hundred baud. L. Hayward proposes a non-return to zero recording system for the Nascom 1 and 2 – the circuit should be adaptable to others – and compares performance with that of the Kansas City interface.

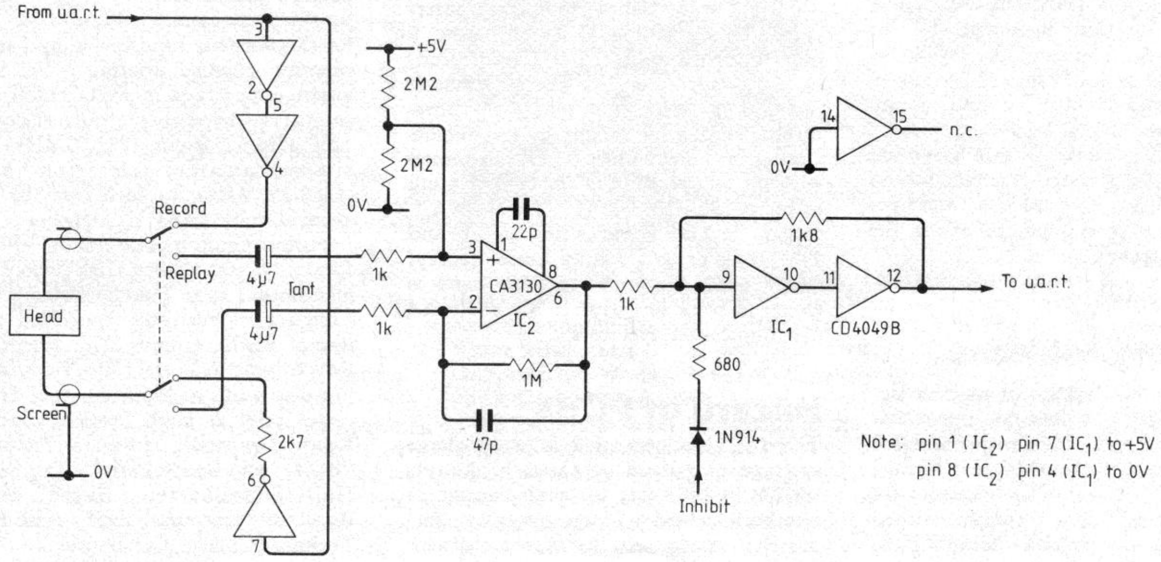
By L. Hayward

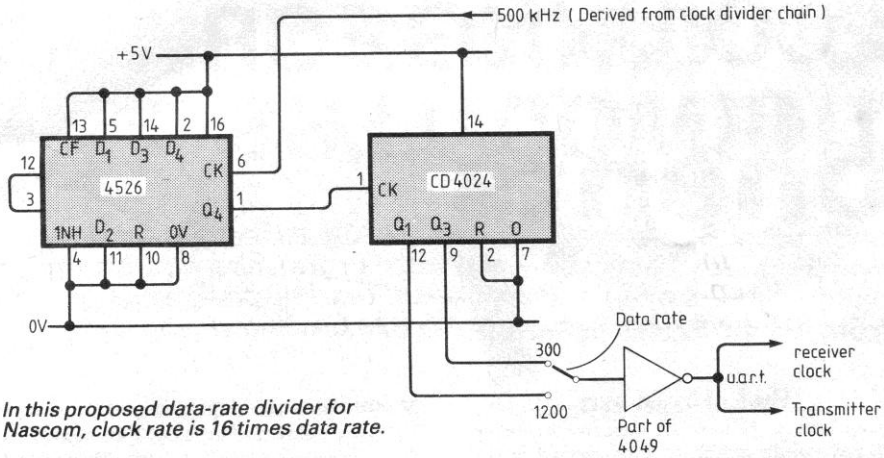
Most small computer systems, in particular those machines offered to the amateur user, have adopted the audio cassette as a convenient method of data storage. Cassette players are readily available at a low price, the only reasonable alternatives being open-reel recording or expensive disc drives. The Kansas City recording standard, developed to use the audio cassette, works well and has become popular due to its tolerance of tape-speed variation, typically 30%. This allows users to exchange recordings between machines of almost any type. But this speed tolerance is the only significant advantage of the system – its disadvantages are susceptibility to tape dropout and slow data speed. Some systems optimistically offer a data rate of 1200 baud, but using cheap cassette decks the best that can normally be attained is 300 baud. The encoding and decoding circuitry involved is fairly complex, using as many as seven or eight large-scale integrated circuits, and enables the user to adapt either the existing audio cassette recorder, or use a "bare-bones" deck, with mechanism and record-playback head only.

Using one of the worst cassette decks I have encountered reliable recording was achieved to a rate of 1200 baud, with fair operation to 2400 baud 1⅞ in/s. The only disadvantage of this is its in tolerance of speed variation: 5% instead of the 30% offered by Kansas City, assuming that the usual uart (universal asychronous receiver-transmitter) is used in the computer. This speed restricition will normally only be a disadvantage if tape transfers from one recorder to another are to be made. Any recorder having a cyclic speed variation of greater than 0.5% would be totally useless for musical reproduction, and most cassette recorder mechanisms can achieve better speed regulation.

The n.r.z. system is well known, and has been used in computing systems for years. No h.f. head bias is used, and the tape is magnetically saturated in a negative or positive sense, depending on whether a zero or a one is written. There is no condition of zero flux, hence the name: non-return to zero. As the tape is saturated no erase head is required, and the system is less sensitive to tape drop-out or variation between various types of tape. Ideally, the system should use heads and tape designed specifically for this type of operation; practical results have shown however, that ordinary heads and tape are quite suitable. The use of certified digital cassettes such as the Scotch type 834A is recommended for the most reliable operation. The circuit shown was specifically designed for use with Nascom 1 and 2 computers, but







In this proposed data-rate divider for Nascom, clock rate is 16 times data rate.

should be suitable for most other computer systems with little change.

Principles of operation

Record mode. The data output from the u.a.r.t. is applied to a 4049 buffer. Sections of this buffer are connected so that the current through the head and current-limiting resistor is reversed in direction as the input changes from logic high or low. The choice of resistor shown is suitable for a typical cassette recorder audio head. It may be reduced if more current is required, up to a maximum of 10 mA for other heads.

Replay mode. A 3130 op-amp is used as a differential amplifier. This mode of operation permits considerable hum to be picked up in the head connecting wires without interference. The gain of 1000 was found to be sufficient to cause the output to clip when replaying data. The input coupling capacitors were made much larger than necessary, considering the frequencies involved, to permit the head to short-out the amplifier input at low frequencies and prevent hum pick-up by capacitive stray coupling. The high frequency response of the amplifier is rolled off to avoid possible pick-up from the nearby clock generator and dividers in the computer.

As the voltage output from the head is proportional to the rate of change of flux the amplifier output will consist of narrow pulses coincident with the timing and direction of the data. In between these pulses, the amplifier output falls to 2.5V. A Schmitt trigger circuit using part of the 4049 is used to hold the state of the previous positive or negative excursion, and thus output restored data to the u.a.r.t. Hysteresis is used to make the output insensitive to spurious small outputs from IC2. The u.a.r.t. requires that the receiver input terminal remains high until the data transmission begins. An inhibit input is provided, which when high prevents IC2 from changing the Schmitt trigger output. This point is conveniently connected to the drive l.e.d. transistor collector in the Nascom, thus making the computer ignore all data until the 'c Load' or 'R' command is executed. The suggested divider circuit is useful if the standard data rates of 300 and 1200 baud are required from the Nascom 1. Power supply required is a single +5V supply; current drain is so small that an existing computer supply should easily accommodate it.

I suggest that circuits such as this be included in small computer systems as an alternative or addition to Kansas City. It shouldn't be too difficult for manufacturers of ready-built systems to offer a completed cassette system as part of the package. If such devices are made available with accurate speed control, thus giving interchangeability, it is likely that the more logical n.r.z. will be adopted universally.

It should be fairly easy to produce a machine with a speed correct to within 5% for reasonable cost. A normal diesel engine with a crude mechanical governor can meet 5% regulation of speed, so why not a simple cassette drive? □