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Commentary

A FEW months ago, last August to be precise, we made reference to the ever-increasing flow of scientists and technicians from this country. This efflux of valuable man-power has been accelerated by recent events and there is no doubt but that it constitutes a real threat to our future economic stability.

There is, of course, a world shortage of technical man-power and this is especially apparent in the U.S.A. and Canada; the two countries which are making the most impact on our reserves of scientists and engineers. In America the shortage is regarded as a serious threat to their national security and economic progress. The competition for the services of qualified engineers is so keen that starting salaries have been forced up some 5 to 10 per cent compared with last year and an engineer graduating this year can expect to earn some $450 a month as soon as he commences industrial employment. At the present rate of exchange this is equivalent to some £1 900 per annum and, although the official rate of exchange may not give a completely realistic comparison, there is no doubt it is considerably more than a young engineer can expect to receive in his first year out of university in this country. A similar position holds in Canada where industry is expanding rapidly and it is recognized that it is essential to import technical manpower in considerable quantity. The recently issued preliminary report of the Gordon Commission on Canada's economic prospects states that during the next twenty-five years the net earnings of the workers may increase by about 75 per cent—a forecast which makes immigration very attractive indeed to any young person.

While we may sympathize with other countries, and especially those of the Commonwealth, on their labour shortages, our own plight is probably worse than any, for, if we are to raise, or even maintain, our present standard of living, we must export on a very large scale. We have few, if any, raw commodities in exportable quantities and virtually the only things we have to sell abroad are skill, craftsmanship and production 'know-how'; a fact that makes it imperative that we should build up and retain a large and efficient force of scientists, technologists and craftsmen of all grades.

As stated previously, this shortage of technical manpower is a worldwide problem and all the major countries are tackling it with considerable rigour. There is no short-term or quick solution to the problem and most countries are tackling it in basically the same manner; by incentive and education—by providing incentive to go in for a technical career and by providing the right type of educational facilities in the appropriate quantity. We, in this country, need a further stage of incentive—incentive for engineers and craftsmen to remain in this country after they have qualified for, as pointed out in our previous editorial, the finest schemes of education and training are of little avail if the best products of such schemes are to be enticed away.

The provision of adequate educational facilities has received considerable attention in the last year or so and the plans made should go a long way, but by no means all the way, towards providing us with a sufficient number of qualified engineers and scientists in future years. The supply of a sufficient number of teachers of the right quality is probably the greatest problem and this applies to teaching at all levels, for the work of the universities and technical colleges is obviously eased if their intake has been well prepared at the grammar schools, and so on down the scale. Here again, we cannot get far without bringing in the word incentive, for to obtain the right type of person, teaching as a career must be made at least as attractive as a career in either industry or commerce.

The incentives required to encourage young people to go in for a technical career and to stay in this country are much the same. The financial reward must be adequate and the conditions of work and the prospects of promotion must be good. In respect of the first point, it must be recognized that, other things being equal, there is a maximum amount which can economically be paid for any particular task, but there is here, of course, another aspect which was brought out forcibly by one of our industrial leaders, Sir Thomas Sopwith who, in a recent company speech said, "We are losing too many of our skilled technicians and craftsmen to countries whose more enlightened fiscal policies enable the individual to retain for himself a much larger share of the rewards accruing to him after very long years of study and effort".

With regard to the prospects of promotion, this is largely a matter for individual firms, who can do much to maintain the enthusiasm of their senior staffs by making it clear that they can progress beyond the laboratory or the factory floor and attain the highest executive level—providing, of course, that they fit themselves for these positions. It may perhaps be remarked that many firms could benefit considerably by having men with sound technical knowledge and experience in their boardrooms.

A Valve-Voltmeter for Synchro Testing

By D. L. Davies*, B.Sc., A.M.I.E.E.

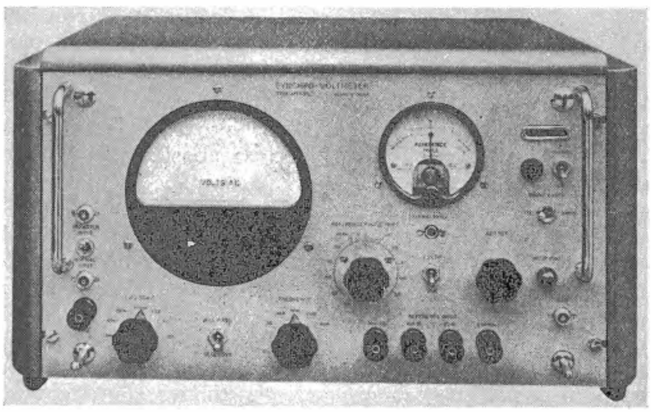
Recent developments in data transmission and allied fields have led to considerable improvements in the performance of all types of synchros. It is necessary, when testing these instruments, to accurately detect null positions, and to examine the residual signal at these points.

The instrument described has been designed to facilitate these measurements. It consists of a phase-conscious meter, and phase-shifter, for accurate detection of a null in the reference component of the test signal; an accurate wide-band valve-voltmeter for measurement of total residual signal; and a selective amplifier to separate quadrature fundamental from the harmonics and noise.

THE testing of synchros involves the detection and examination of null voltages produced under test conditions[1].

It is normally required to detect a null in that component of the test signal which is in phase with a reference source; usually the carrier supply to the rotor of the synchro concerned. When a null has been achieved, it is then required to measure the total residual signal at that point. This residual signal consists of a component at carrier frequency but of quadrature phase (since the reference component has been made zero in the nulling process), harmonics of this frequency, and of sundry other noise components. It is usually required to isolate and measure the quadrature component of carrier frequency.

The instrument described was developed to fulfil these objectives.



The complete instrument

In order to fulfil its first function, i.e., to detect a null in the reference component of the test signal, the instrument contains both a phase-conscious meter indicator, and a phase-shifting network. The phase-conscious centre-zero meter indicates when the reference component of the test signal passes through zero, and the phase-shifting network introduces a variable known phase-shift to compensate for the phase shift through the synchro windings at carrier frequency.

To satisfy the second function, i.e., to measure the residual signal at a null point and isolate the fundamental frequency component of it, there is a valve-voltmeter of the gated feedback type which can be made selective by the inclusion of bandpass filters in the signal path.

The instrument can be broken down into functional blocks, which are described in the following sections.

Circuit Arrangement (Fig. 1)

Referring to the block diagram, it is seen that the test signal can be applied directly to the input amplifier ('Normal'), or via a 10:1 attenuator. The purpose of this amplifier is to act as a buffer between the signal input source and the comparatively low impedance range attenuator, i.e. it has a stable gain of unity, a very high input impedance (greater than 30MΩ in parallel with less than 20pF), and a very low output impedance.

The signal then passes through the range attenuator, which is the sensitivity control for the whole instrument.

A stage of gain follows (amplifier *A*), to bring the signal up to a more convenient level for application to the filter networks. By means of a 'Wide-Band/Selective' switch, the signal is then routed either through a direct channel to the main meter, or via filter circuits, before returning to the main signal channel.

In the 'Wide-Band' position, further gain is applied to the signal before it reaches the output amplifier and meter circuit (amplifier *B*). Amplifier *B* incorporates variable resistors in its feedback path which are ganged to the main sensitivity switch, thus enabling the overall gain from input to meter to be accurately adjusted for each position of the range attenuator. The output of this amplifier is made available to the user for monitoring purposes (c.r.o. socket). It is operative in both the 'Wide-Band' and 'Selective' conditions.

The output amplifier and meter circuit have a form of feedback applied over them, sometimes referred to as 'gated feedback' since it involves the use of biased-off diodes which are 'gated' on and off by the signal applied, completing the feedback loop only when they conduct. The purpose of biasing off the diodes is to obviate the standing current which is present in more conventional circuits using vacuum diode rectifiers. Since vacuum diodes are used, their back impedance can be virtually ignored, and when conducting they include the meter in a tight feedback loop, thus eliminating the effect of any variations in their forward characteristics. Used in the 'Wide-Band' condition, the main meter has a linear response at all sensitivities, from 15mV to 150V f.s.d., with an accuracy of ± 2 per cent of full scale over the frequency band from 20c/s to 20kc/s.

In the 'Selective' position, the signal is extracted after amplifier *A* and fed via two filter circuits back to amplifier *B*, and hence to the main meter. The two filter circuits are stagger-tuned to give an overall bandpass characteristic. This is done so that variations in the frequency of the carrier supply can be accommodated. Each filter consists of a stable feedback amplifier, with a 'twin-T' network in a feedback path; the gain being adjusted in each case to give an effective Q of 10. Five switched frequencies of operation are provided, being the carrier frequencies in use at the present time, e.g., 60c/s, 400c/s, 1·1kc/s, 1·6kc/s and 2·4kc/s. The overall gain via both filters is adjusted to be

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unity at mid-band, so that the signal returns to amplifier B at the same level as it left amplifier A. The off-tune rejection is such that the response to second harmonics is 75 times down, and that to third harmonics is 150 times down.

The output of amplifier A also forms a convenient point for taking off the signal for use in the phase-shift and phase-conscious meter circuits. The signal must be sampled prior to the filter networks, for they would introduce undesirable phase shifts in themselves, leading to a completely erroneous indication on the phase-conscious meter.

After further amplification (Signal Amplifier), the signal is applied to a well-known phase-shifting network of the constant output-amplitude variety. This phase-shifting circuit produces a variable phase shift centred about 90°, the reason for which will now be explained.

It was desired to produce a phase-shifting network which conscious meter circuit as gating waveforms; i.e., the signal from the buffer cathode-follower is gated through the centre-zero meter at pre-set portions of the reference wave.

A reference level setting device is incorporated, since without it the squaring stages would produce waveforms whose mark-to-space ratio would vary with the amplitude of the carrier supplied, thus causing wrong operation of the phase-conscious meter circuit. In normal use, the reference input level control is adjusted in the 'Set-up' position to produce a fixed deflexion on the centre-zero meter.

The phase-conscious meter circuit will indicate when the applied signal passes through zero phase shift with respect to the reference input, and will ignore to a large degree the presence of quadrature signal, or of harmonic and spurious frequencies.

Main h.t. supplies for the circuits are provided via a self-

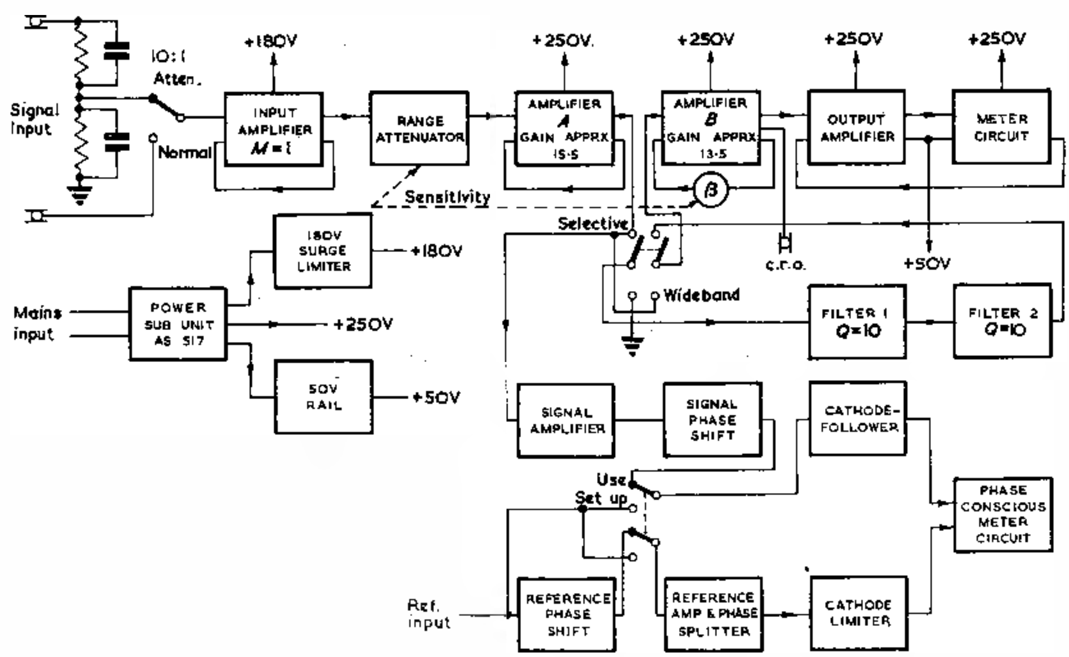


Fig. 1. Circuit arrangement

would produce an effective phase-shift in the reference channel of $\pm 60°$. Because of the difficulty in devising a circuit which will vary phase smoothly through zero, with no amplitude variation, it was decided to phase-shift both the reference and signal channels by 90°, and then to vary the amount in each channel about the 90° point, but in opposite directions; thus producing a phase-shift between the channels. It is comparatively easy to produce a constant output-amplitude phase-shifter which varies in either direction about 90°, and hence it was possible to produce a phase-shift between the channels which could be varied in either side of, and through, zero phase shift. A ganged control was used for the two channels, calibrated on the front panel in terms of the effective phase-shift in the reference channel.

Having undergone some phase-shift, the signal is then fed via a buffer cathode-follower to the phase-conscious meter circuit.

The reference input is fed via a level-setting device to the reference phase-shift network, which is identical with that in the signal channel. After phase-shift, the reference voltage undergoes further amplification before being applied to a phase-splitter which produces two outputs in anti-phase. These two outputs are each fed to squaring stages of controlled mark-to-space ratio, the resultant rectangular waveforms being then applied to the phase-contained stabilized power supply sub-unit, delivering 250V, but a 50V supply is derived from it by means of a sampling chain and a low output impedance amplifier. Trouble is often experienced with high sensitivity indicators due to mains 'bounce', so that in this instrument it was considered desirable to embody a surge limiting stage which acted as a buffer between the low level input amplifier and the main h.t. line.

Circuit Details

Input Attenuator (Fig. 2)

This is a simple resistive attenuator, consisting of R_1 in the top arm, with R_2, R_3 and VR_1 in the bottom arm. VR_1 is used for precise adjustment of the attenuation ratio. C_2 and C_3 provide high frequency compensation, C_2 being adjustable for optimum response. The input is blocked via C_1.

Input Amplifier (Fig. 2)

This consists of two cascaded gain stages V_1 and V_2, d.c. coupled to a cathode-follower V_{3a}, the output of this latter stage being fed back in its entirety to the cathode of V_1, giving an overall gain of very nearly unity. This circuit arrangement is often termed a 'ring of three', for obvious reasons. C_5, C_7, C_8 and R_{12} are included to ensure loop stability at high frequencies.

RANGE ATTENUATOR (Fig. 3)

The blocking capacitor C_{11} feeds the signal from the cathode of V_{3a} to the range attenuator. This consists of 1 per cent high stability resistors, some odd values being achieved by paralleling. The total resistance of the attenuation is such that it does not require capacitive compensation over the range of frequencies for which it is intended.

AMPLIFIER *A* (Fig. 3)

The grid of V_{4b} is fed directly from the switched output of the range attenuator, and its anode is a.c. coupled to the grid of V_5; the output from this stage being d.c. coupled to the cathode-follower V_{4a}. The signal at this point is then fed back to the cathode of V_{4b} via the attenuator formed by R_{40} and R_{41}. This circuit is another example of the 'ring of three' technique. As in the previous example, the second interstage coupling is a direct one in the interests of loop stability at low frequencies. C_{15}, C_{13} and C_{31} are again included for loop stability reasons at high frequencies. The gain of this stage is roughly given by $(R_{40} + R_{41})/R_{41}$, i.e., just over 15 times.

AMPLIFIER *B* (Fig. 4)

While the general form of this amplifier is identical with amplifier *A*, the feedback attenuator contains switched preset components VR_{2-8}. The switch is ganged to the range attenuator switch, so that at each position the nominal gain of about 13 times is capable of some small adjustment. Again, the coupling between V_7 and V_{6b} is direct, while C_{18}, C_{20} and R_{44} are included for stability purposes. The input to this stage is via a blocking capacitor C_{16}.

OUTPUT AMPLIFIER AND METER STAGE (Fig. 5)

Another 'ring of three' circuit, with both halves of V_{10} in parallel as the output cathode-follower; this amplifier drives the meter M_1 via the rectifier V_{11}. Standing bias for the diodes is provided by the d.c. current flowing via R_{71} and R_{72}, and feedback to the cathode of V_8 is taken from across R_{74}. The screen grid of V_8 is fed from the junction of R_{75} and R_{78}, thus providing some decoupling of that electrode. Again, in the interests of loop stability, V_9 is directly coupled to V_{10}, and C_{23} is provided. In common with the previous amplifiers of 'ring of three' form, the top arm of the direct coupling is by-passed to signal frequencies; in this case by C_{22}.

FILTER 1 (Fig. 6)

The input to this stage is provided by the output of amplifier *A*, via the 'Wide-Band/Selective' switch, and is first amplified in V_{12}, the anode of which is directly

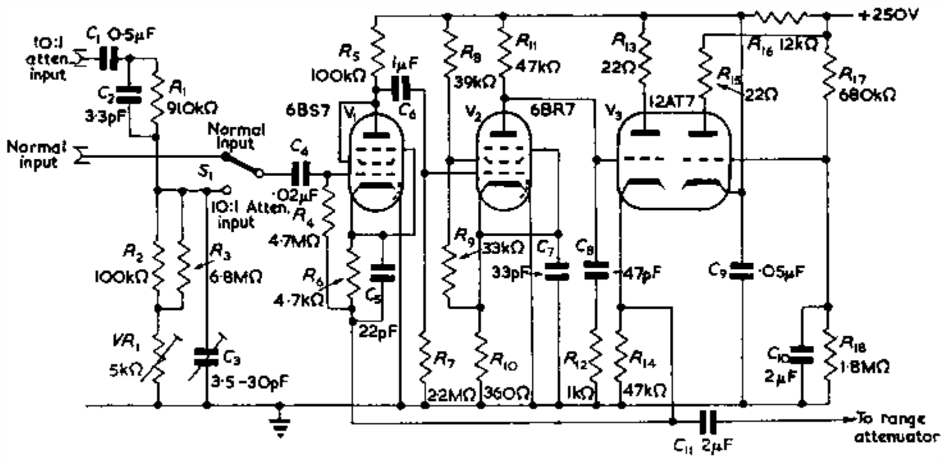


Fig. 2. Input amplifier

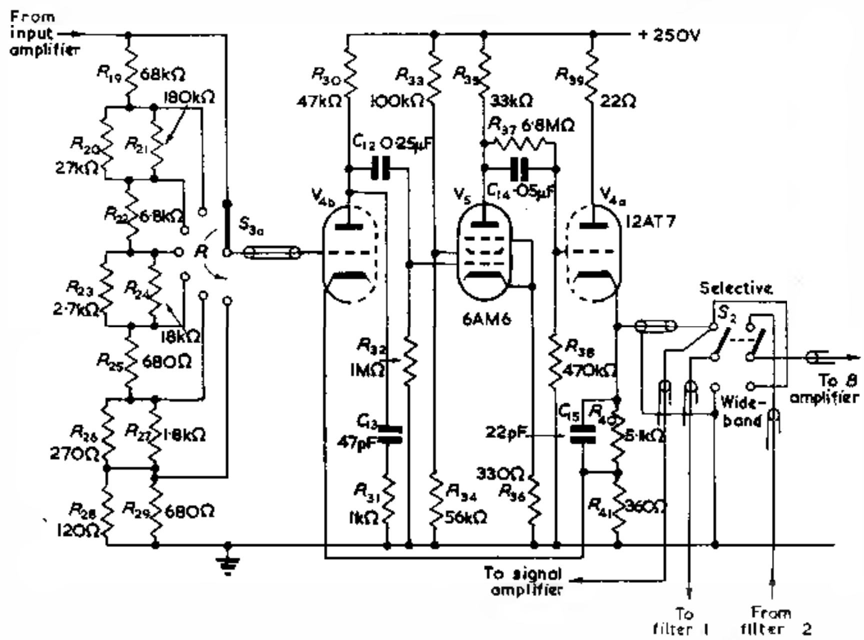
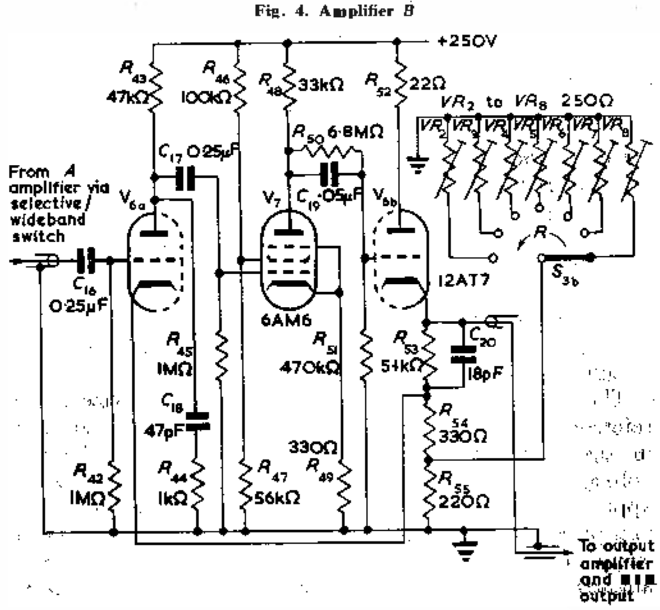


Fig. 3. Range attenuator and amplifier *A*

Fig. 4. Amplifier *B*



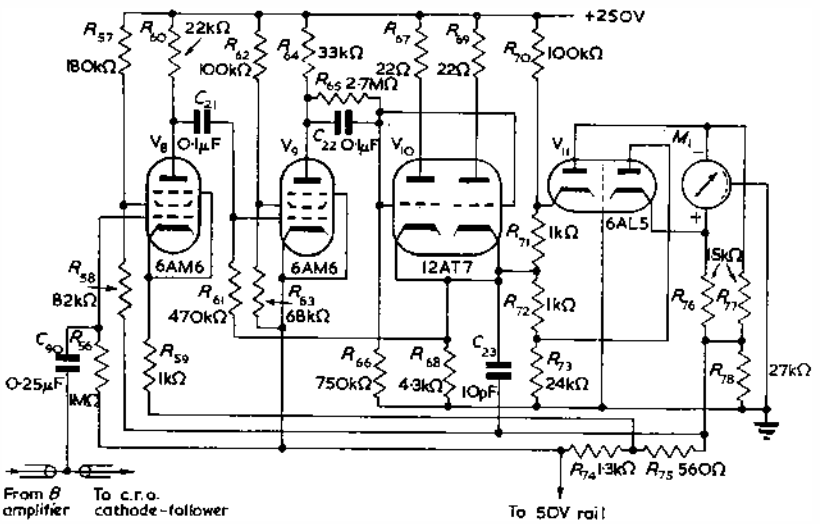


Fig. 5. Output amplifier and meter circuit

coupled into the cathode-follower V_{13a}. This output cathode then drives the 'twin-T' selective network, and also a resistive attenuator consisting of $VR_{9\text{-}13}$, R_{90} and R_{91}, which is returned to the output of the cathode-follower V_{13b}, the grid of which is fed from the output of the selective network. The junction of R_{91} and R_{90} is taken to the cathode of the input valve V_{12a}.

Consider the action of the circuit at the tuned frequency of the selective network. Due to the null transmission of the 'twin-T', there will be virtually zero signal at the grid of V_{13b}, and hence also at its cathode. This point can then be considered as a virtual earth return for the attenuator $VR_{9\text{-}13}$, R_{90} and R_{91}. V_{12} and V_{13a} can then be seen to form a 'ring of three', with a feedback attenuator from the cathode of V_{13a} to the cathode of V_{12a}. The gain will be approximately set by $(VR_{9\text{-}13}+R_{90}+R_{91})/R_{91}$ and can be varied by the pre-set potentiometers.

Now consider the action of the circuit at a frequency well removed from the tuned frequency of the 'twin-T', i.e., where practically no attenuation occurs. Then the cathodes of V_{13a} and V_{13b} will be virtually connected together and to the cathode of V_{12a}. Thus the circuit looks like a 'ring of three' with all the output fed back to the first cathode, i.e., with a gain of near unity.

It can be seen that the transmission of the whole filter circuit is a maximum at the tune frequency of the 'twin-T' network, falling off appreciably on either side. It can be shown that the Q of the selective amplifier so formed is determined by the relationship $Q = (M+1)/4$, where M is the gain operative around the loop. In this case, the Q is set to 10, by fixing the gain from the input grid of V_{12a} to the output cathode of V_{13a} at 39 times. The elements of the 'twin-T' are switched, and the gain can be accurately set to $\times 39$ at each switch position by means of the pre-set potentiometers $VR_{9\text{-}13}$.

C_{25} and R_{84} are present for loop stability reasons. The output from the complete filter circuit is taken from the slider of the pre-set potentiometer VR_{27}, which is adjusted so that the overall transmission at tune is $\sqrt{2}$.

Filter 2

This circuit is identical in form with Filter 1, but differs in the values of the 'twin-T' components. These are chosen such that the responses due to the two filter circuits cross over at their 3dB down points, thus giving a flat-topped bandpass characteristic. Since both circuits have a phase-shift of 45° at this point and a transmission of 0·707, then the overall transmission via both filters over the pass band is unity.

The output of Filter 2 is fed via the 'Wide-Band/Selective' switch to the input of amplifier *B*.

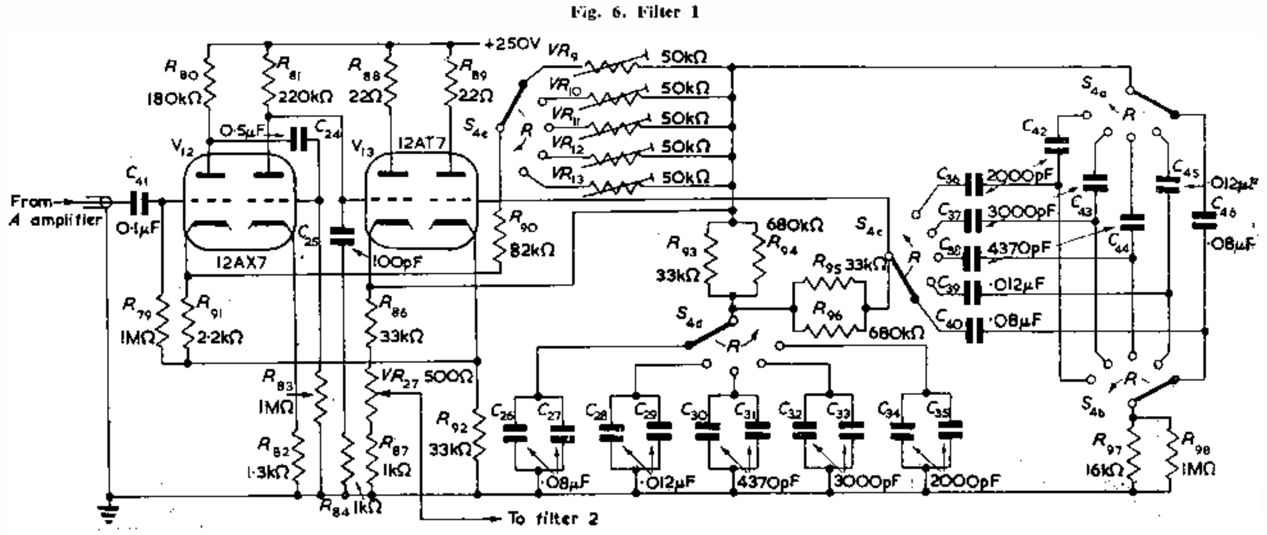
Signal Amplifier (Fig. 7)

This circuit also derives its input from the output of amplifier *A*, via the 'Wide-Band/Selective' switch. It is again a 'ring of three' and, in common with the other examples, it has direct coupling from V_{19} to V_{18b}, by-passed to signal frequencies by C_{78}. C_{76} and R_{138} are included for high frequency loop stability. The gain of the stage is given approximately by $(R_{146}+R_{147})/R_{147}$.

Signal Phase-Shift (Fig. 7)

An output is taken from the anode of V_{18b} in anti-phase to that at its cathode. These 'push-pull' signals are then applied to the network formed by R_{150}, VR_{39}, VR_{21b} and $C_{79\text{-}83}$. An output is taken from the junction of the capacitive and resistive elements. This phase-shift network is

Fig. 6. Filter 1



well known, and gives an output of virtually constant amplitude for wide variations in its C and R values. In this case, the switch which selects the capacitors $C_{79\text{-}88}$ is ganged to the filter frequency selector switch. The variable resistor VR_{21b} is the phase-shift control on the front panel, while VR_{29} is used for setting the limits of phase swing obtainable.

Reference Input (Fig. 8)

The reference supply is connected to a resistive attenuator consisting of R_{152}, R_{153}, R_{154} and VR_{20}, the latter control being marked as 'Set Ref' on the front panel. Its purpose is to adjust the level of the reference voltage within the instrument to be a constant fixed level, irrespective of the voltage applied. Any voltage from 10 to 300V can be accommodated.

Reference Phase-Shift (Fig. 8)

The output from VR_{20} is fed via the blocking capacitor C_{84} to a 'concertina' type of phase splitter consisting of V_{20b} with equal anode and cathode loads R_{160} and R_{157}. The resulting 'push-pull' voltages are then fed to the network formed by R_{158}, VR_{28}, VR_{21a} and $C_{85\text{-}89}$. An output is taken from the junction of the capacitive and resistive elements. The entire circuit is almost identical with the signal phase-shift circuit, but for the fact that VR_{21a} and VR_{21b} are ganged in such a way that, when the value of one increases, the other decreases, and vice versa.

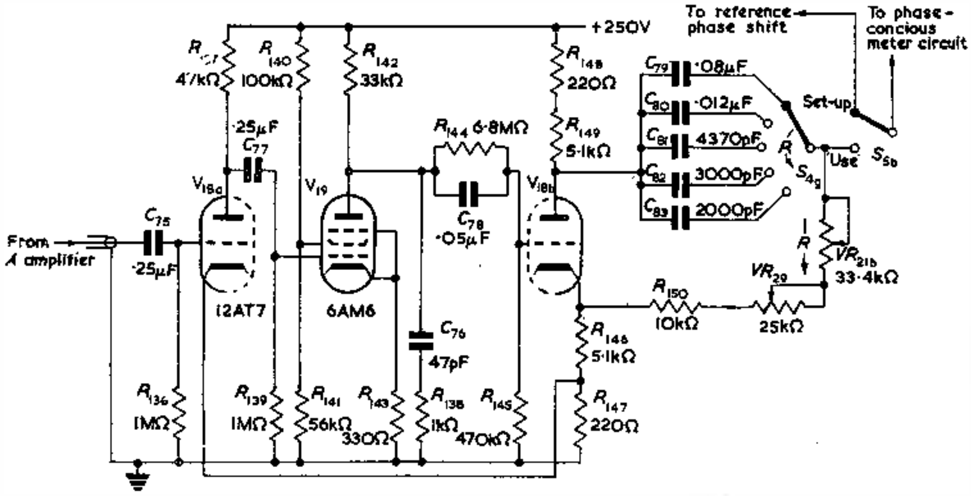


Fig. 7. Signal amplifier and signal phase shift

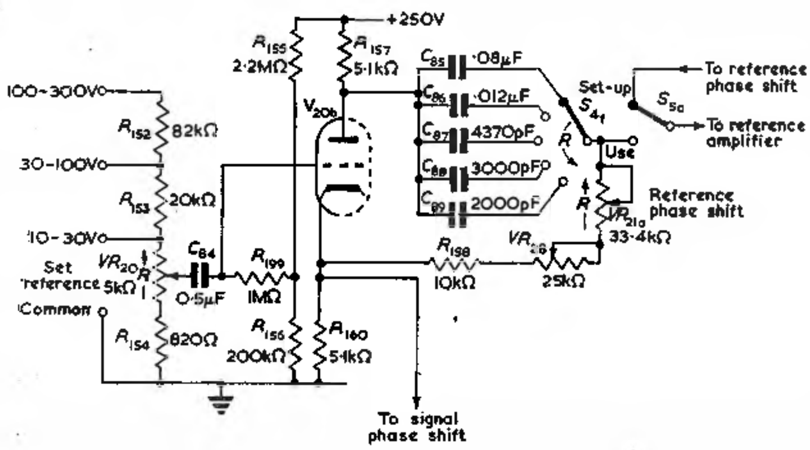


Fig. 8. Reference input and reference phase shift

Reference Amplifier (Fig. 9)

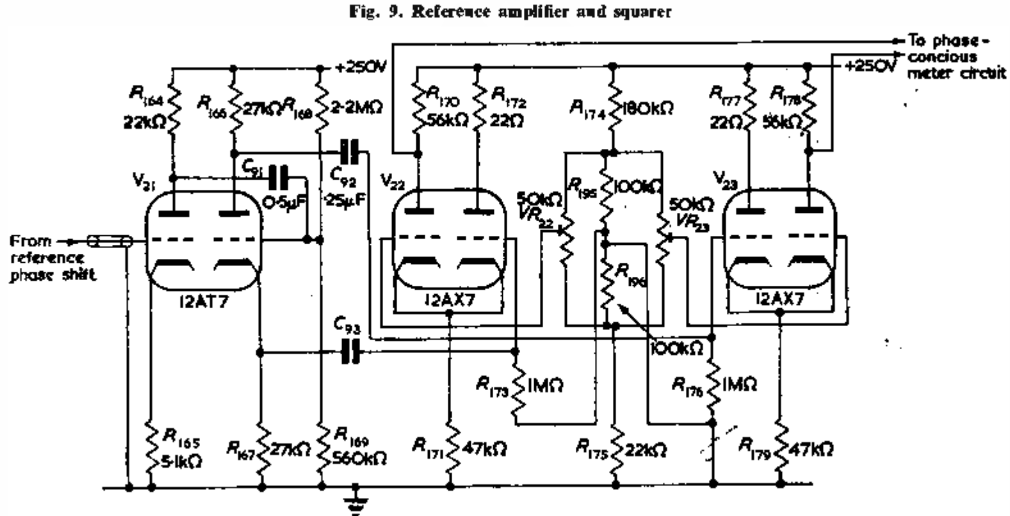
The output from the reference phase shift circuit is fed to the grid of V_{21a} where a gain of about 4 is produced, being set by the ratio of R_{164} to R_{165}. In the 'set-up' position of the 'Set-up/Use' switch, the grid of V_{21a} can be fed from the cathode of V_{20b}, i.e., with a voltage to which no phase-shift has been applied.

The output at the anode of V_{21a} is connected to the grid of V_{21b}, which is another 'concertina' type of phase-splitter, having equal anode and cathode loads R_{166} and R_{167}.

Squarers (Fig. 9)

The 'push-pull' outputs from V_{21b} are each fed to separate squaring stages, formed by V_{22} and V_{23}. These take the form of cathode-coupled limiters. Considering V_{22}, it can be considered as a cathode-follower V_{22b} feeding a grounded grid stage. V_{22a}, with the output taken from the anode resistor R_{170}. The mark-to-space ratio of the resulting rectangular waveform can be controlled by the d.c. level setting potentiometer VR_{22}.

Fig. 9. Reference amplifier and squarer



The circuit of V_{23} is identical with that of V_{23}, resulting in an output from R_{178} of mark-to-space ratio controlled by VR_{23}.

Signal Cathode-Follower (Fig. 10)

This stage, consisting of V_{20a}, acts as a buffer between the output of the signal phase-shift circuit and the phase-conscious meter circuit. Its input can be supplied with a non-phase-shifted voltage from the cathode of V_{20b} when the 'Set-Up/Use' switch is in the 'Set-Up' position. The output feeds the junction point of the cathode resistors R_{185} and R_{190}, this being the point of injection of the signal voltage into the phase-conscious meter circuit.

Phase-Conscious Meter Circuit (Fig. 10)

This circuit has been described elsewhere[2], and so only a very short summary will be given here. It consists of two identical cathode-followers V_{24b} and V_{24a}, their grids being fed with d.c. restored rectangular waveforms of fixed mark-to-space ratio, derived from the squaring stages previously described. MR_1 and MR_2 respectively are the diodes used for d.c. restoration, MR_2 being returned to the decoupled point at the junction of R_{197} and R_{198}, and MR_1 being returned to the decoupler slider of VR_{25}. This allows for any difference in bias potentials of V_{24b} and V_{24a} to be taken up. The centre-zero meter M_2 is connected between the two cathodes, with series resistors R_{187}, R_{188} and VR_{24}. VR_{24} is included so that the meter sensitivity can be adjusted.

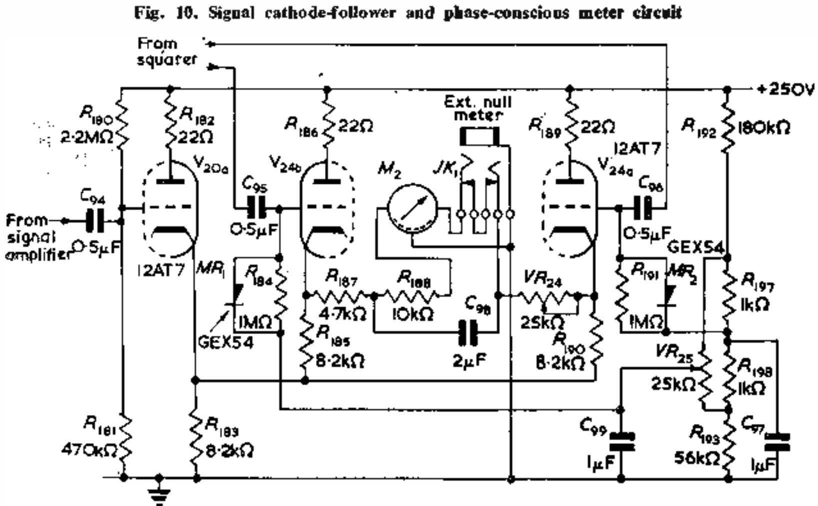
The action of the circuit is such that the rectangular gating waveforms applied to the grids of V_{24} route the signal from the cathode of V_{20a} through M_2 in different directions for different portions of the cycle, thus producing rectification. The mark-to-space ratio of the gating waveforms is chosen for best rejection of harmonics that may be present in the signal waveform. A rejection of approximately 100 : 1 is achieved for both 2nd and 3rd harmonics.

The meter jack JK_1 provides for the use of an external meter in series with M_2.

50V Supply (Fig. 11)

This consists of an adjustable divider chain across the main 250V supply, consisting of R_{119}, R_{120} and VR_{19}. The output is smoothed by C_{70} and applied to the grid of V_{16b} which, together with V_{16a} and V_{17a}, forms a directly coupled 'ring of three', providing a low output impedance, ripple free, source of 50V at the cathode of V_{17a}. C_{73} and R_{131} are provided to ensure high frequency loop stability, while C_{71} and C_{72} increase the loop gain at high frequencies, and hence keep the output impedance low there.

Fig. 10. Signal cathode-follower and phase-conscious meter circuit



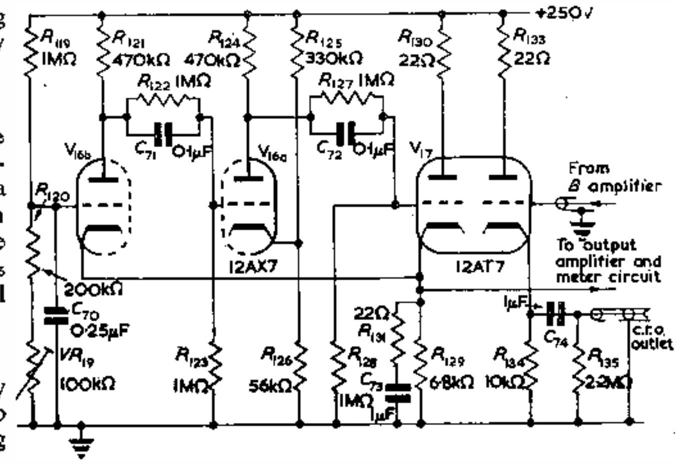


Fig. 11. 50V Supply

C.R.O. Outlet (Fig. 11)

The grid of the cathode-follower V_{17b} is fed from the output of amplifier B. An output is taken from its cathode via the blocking capacitor C_{74} to SK_3 on the front panel. The stage merely acts as a buffer.

Surge Limiter (Fig. 2)

The h.t. supply for the input amplifier comes partly through the bleed resistor R_{16} and partly through the cathode-follower V_{3b}. The grid of this valve is held at about 180V by the divider chain formed by R_{17} and R_{18} across the 250V supply. This point is smoothed by C_{10}, and the cathode of V_3 is by-passed to ground by C_9. Any surges that occur on the main 250V supply are greatly attenuated by the combination of R_{17}, R_{18} and C_{10}, and consequently so is the output at the cathode of V_{3b}.

Earthing

A switch is provided on the front panel so that the chassis of the instrument can be either connected directly to mains earth, or via a 47Ω resistor, the intention being to reduce spurious voltages produced by possible earth loops.

Conclusions

At the time of writing, upwards of two dozen instruments were in regular use for purposes of synchro performance testing, either by manufacturers or by large-scale users of synchros. Active use over several months, and with a wide variety of test arrangements, had produced no major criticisms. It would appear that the instrument described does, in fact, successfully perform all the functions for which it was originally designed.

Acknowledgments

The author wishes to record his thanks to the Directors of Solartron Research & Development, Ltd for permission to publish the details of the instrument described.

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An Accumulator Unit for a Dekatron Calculator

By R. Townsend*, M.Sc., A.M.I.E.E., and K. Camm†, B.Sc., A.C.G.I.

This article describes a Dekatron unit in which decimal numbers may be added or subtracted; the sum being returned to the main store multiplied or divided by ten or unity.

The calculator, for which this was designed, uses cold-cathode trigger tubes and Dekatrons as its main elements, operates in a parallel mode and takes its input from punched cards. The 'carries' between columns are added in one carry cycle by the use of 'nine-sensing' circuits.

THE unit had to facilitate several operations which were basically required for the calculator.

It had to be capable of:

(1) Being cleared of any previously existing number, leaving zero in all columns.

(2) Taking numbers from temporary storage addresses adding and/or subtracting them from the cleared accumulator, or from any number already present in the accumulator.

(3) Returning a number to any selected storage address.

(4) Multiplying or dividing a number by ten, by simply shifting all the digits one column to the left (multiplying) or one column to the right (dividing).

Data Flow

The flow diagram of Fig. 1 shows the accumulator linked to the storage addresses via the highways. Each column of the accumulator is permanently connected to an input and an output highway; the names 'input' and 'output' being referred to the store. For convenience in the diagram, the accumulator is shown as having only four columns, whereas it would have up to fourteen. Likewise, each storage address is shown as having only four columns, whereas each address may be made up, as required for the program, in convenient numbers (2, 4 or more) of columns, being plugged column for column to the highways.

Only those addresses required for use by the program would be plugged to the highways. In the figure, three storage addresses are shown plugged.

Only one storage address is involved with the accumulator per calculator cycle, either reading into or out of the accumulator. The address selector mesh gates pulses to one only of the plugged addresses per calculator cycle.

In general, numbers are sent at card reader speed from fields on the punched card into the appropriate storage addresses. From the selected address, the number is then sent at calculator speed on to the output highways and into the accumulator. From the latter, the number is sent at calculator speed on to the input highways and hence into a second selected address. From certain of these addresses the number can be read out at card reader speed to a print unit or card punch.

The techniques of transferring the numbers from storage addresses to accumulator and vice versa, are identical.

Drive Pulses and Calculator Pulse Cycles

Since both the main storage and the accumulator use

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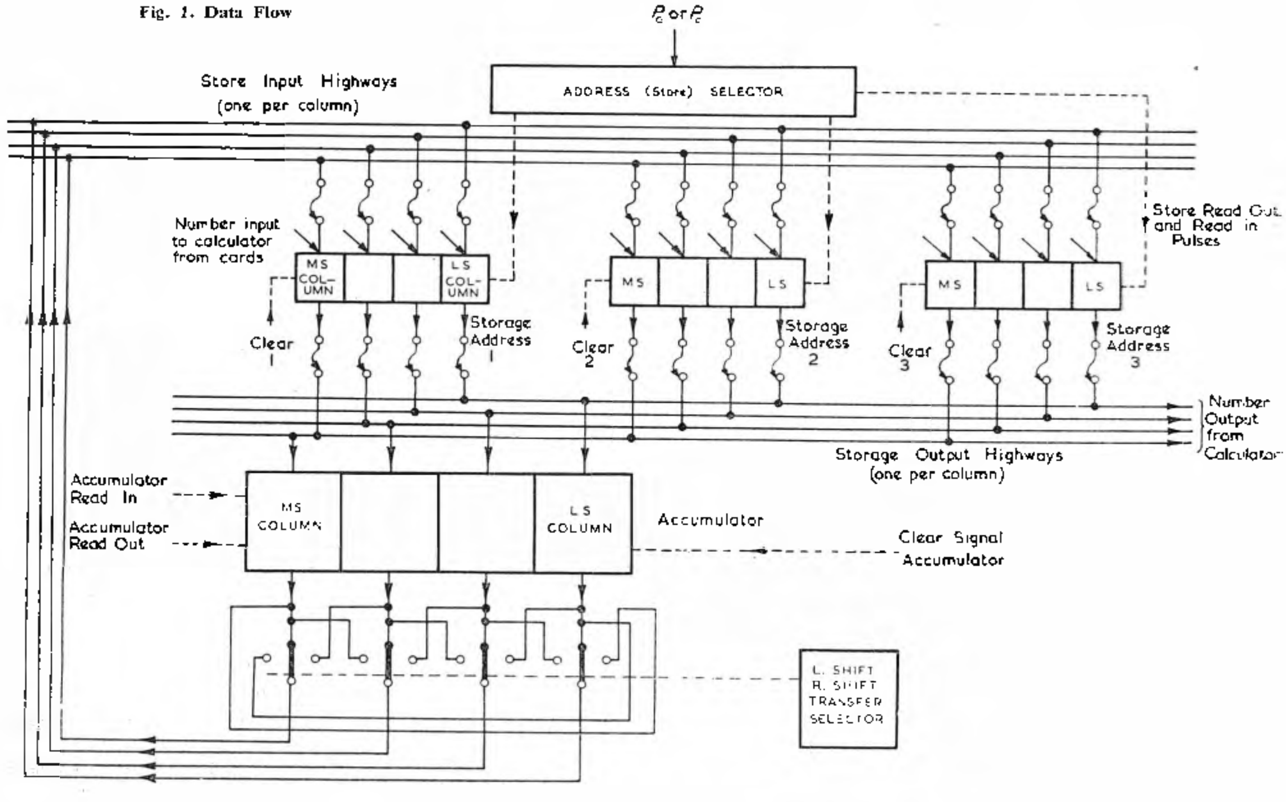
† *Ferranti Ltd, formerly The British Tabulating Machine Co. Ltd.*

Fig. 1. Data Flow



Dekatrons as the storage medium, a brief description is required of the method of driving the Dekatrons, although their operation has been described more fully in previous literature[1].

The Dekatrons in the store and the accumulator are both Ericsson types, having two sets of guides. Two separate and consecutive pulses of 100V amplitude are required; one to the first guides and the other to the second guides, for shifting the discharge glow from one cathode to the next.

Following normal practice, the calculator generates clock pulses at 100kp/s and produces from them a 5-pulse minor cycle, as shown in Fig. 2. Each pulse in the minor cycle has a recurrence frequency of 20kp/s and a nominal duration of 10μsec. Pulses P_a and P_b require widening by 10 to 20 per cent to provide the time overlap required for double-pulse operation of the Dekatrons[2].

A unit step on the Dekatrons is carried out once in a

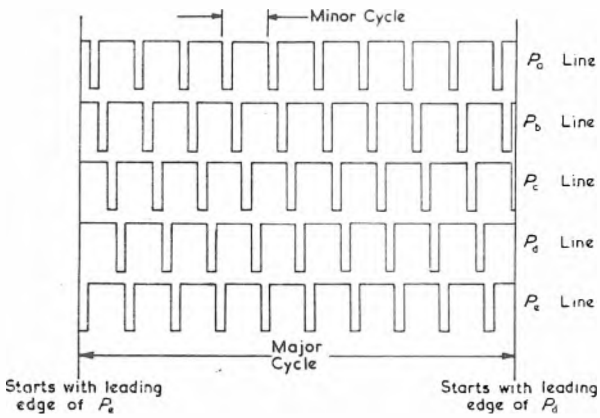


Fig. 2. Pulse cycles used in calculator

minor cycle. By taking successive P_a, P_b and P_c pulses from their individual lines and always supplying the middle one (P_b) to the Dekatron second guides, the direction of rotation of the discharge glow around the tube is determined by the application of either P_a or P_c to the first guides, as shown in Fig. 3. The two remaining pulses of the minor cycle, P_d and P_e, are used for gating.

Minor cycles are used in the calculator in groups of ten, known as a major cycle. One major cycle is gated out per control pulse from the calculator control, this being sent after the address mesh has selected the address, and the order register has set up the pulse distribution paths required by the next order.

A major cycle is the basis of an addition, subtraction, transfer, left shift or right shift; and by programing, multiplication and division.

The Transfer of Numbers from Store to Accumulator (Read In)

Fig. 4 shows one column of a storage address and its plugged connexion via the highways to the corresponding column of the accumulator.

Whatever the function, the discharge glows in the accumulator always rotate clockwise. The discharge glows in the store Dekatrons rotate clockwise if it is desired to read the stored number into the accumulator as a true value for addition, or anti-clockwise if the number is required in the complementary form for subtraction.

To read into the accumulator, for either addition or subtraction, the order mesh sets up:

(1) A path to the accumulator read-in terminals to feed P_a and P_b pulses when the major cycle is called, feeding either (a) the whole major cycle for addition, or (b) a modified major cycle with the last minor cycle missing, for subtraction.

(2) A path to all the second guide terminals of the plugged addresses to feed ten P_b pulses when the major cycle is called.

The reason for the modified major cycle in the case of subtraction is explained later.

An address is selected, containing a number x which may be either added to or subtracted from the accumulator contents, noting that these initial contents may be required to be zero. Zero-ing, or clearing, can be manual or by program order such as 'clear and add' or 'clear and subtract' in which case a clearing signal is fed to the 'clear accumulator' terminal before the major cycle is called.

This signal is a negative pulse of 100V amplitude which pulls the potential of the '0' cathodes of the accumulator Dekatrons down to −100V by means of the unistor diodes D_1, pulling the glow on to this cathode in each tube.

By means of the address mesh already mentioned, this clearing signal can also be applied selectively to any storage address.

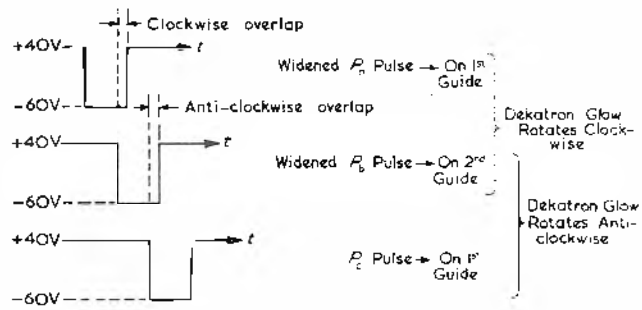


Fig. 3. Three-pulse system for reversible Dekatron drive

Selection of an address to be used with the accumulator is achieved by the address mesh setting up a path to the selector terminal of that address to feed major cycle pulses.

The order mesh decides whether these pulses shall be P_a or P_c; P_a for addition, P_c for subtraction.

Addition

The paths are set up by the address mesh and the pulses are selected by the order mesh, before the control calls for a major cycle. During this cycle, ten P_a and ten P_b pulses are received by all the Dekatrons in the selected address, stepping each glow once around the tube face, leaving the original number still stored in the storage address.

Referring to Fig. 4, if x is the digit stored in any column n of the address, then after $10-x$ minor cycles the glow arrives at the '0' cathode raising both its potential and, by capacitive coupling, that of the trigger electrode of the G1/371K trigger tube, V_2. The anode of this trigger tube has been plugged to the column highway, which also has permanently connected to it, the input to that column in the accumulator. When the trigger tube is 'off', the highway potential is at +40V (h.t.); when the tube is struck the highway potential is drawn down to −60V. The accumulator input from each highway controls a diode gate in the identified column. When the highway is at +40V, current flows through the unistor diode Q_1 and the resistor R_g ($R_g \gg$ the anode load of the trigger tube R_a) so that the junction of the diodes Q_1 and Q_2 is held at approximately +40V despite any negative-going voltage pulses applied to the opposite end of R_g. When the highway is at −60V, the junction of the diodes is allowed to follow the negative pulses on R_g until limiting takes place at −60V. The Dekatron guide potential is lowered by virtue of the

Q_2 diode connexion. The d.c. potential of the highway acts as a gating voltage to any pulses impressed on R_g. The second guide is gated similarly by the same potential. The guides of an accumulator column Dekatron will not receive any driving pulses until the trigger tube in that column of the selected storage address is struck.

The address Dekatron rotates clockwise for addition, and therefore the trigger tube is struck after $10 - x$ pulses in the major cycle have passed. During this time, although P_a and P_b pulses are routed to the accumulator read-in terminals, the diode gate is closed, so the glow in this column has not moved.

When the address trigger tube has struck, the remaining pulses in the major cycle are allowed to step the glow in the accumulator column clockwise. These remaining pulses in the major cycle will be x.

If the number y was previously stored in the accumulator column and $x + y \geqslant 10$, the accumulator glow will be either standing on the '0' cathode, or on some cathode following '0' in the clockwise sense. In either event, the trigger tube, V_4, in this accumulator column will have been 'struck'. This denotes a 'carry' to the next column higher in power.

The treatment of the carries is dealt with later. It is clear that all the trigger tubes must be extinguished at some time prior to the start of a major cycle.

Subtraction

The process for subtraction is identical, with the exception that, (1) P_c pulses are sent to the selected storage address to step this anti-clockwise, and (2) a modified major cycle of P_a and P_b pulses is sent to the accumulator read-in terminals.

Both of these features result in the nine-complement of x being added into the accumulator column.

This is necessary if an 'end-around' carry system is used; i.e. each carry feeds into the next column in ascending power, and the highest power column feeds its carry back into the lowest, forming a closed-carry loop.

An example will illustrate this:

Subtract 1272 from 2274 (=1002)

(1) Take the nine-complement of 1272 (which is the remainder when each digit is subtracted from 9) (=8727)

(2) Add this to 2274, taking note of all carries.

$\left.\begin{array}{l}2274\\8727\end{array}\right\}$ requires one major cycle

0991 number in accumulator at end of major cycle (Primary sum)
1001 carries stored on trigger tubes

(3) Shifting carries one column higher around carry loop each time, add them to the sum until no further carries ensue.

0991 = Primary sum
0011 = Shifted carries

0902 ⟶ 0902
0010 ⟶ L Shift ⟶ 0100

0002 — ⟶ 0002
0100 → L Shift → 1000

Final sum 1002

No further carries 0000

This is accomplished by having the address Dekatrons step anti-clockwise, and if x has been stored, the glow will complete one revolution of the tube face, still leaving x stored, but striking the trigger tube after x pulses of the major cycle. During this time, the accumulator column Dekatron remains unchanged, responding when the trigger tube is struck, to the remaining pulses of the *modified* major cycle, which are $9 - x$.

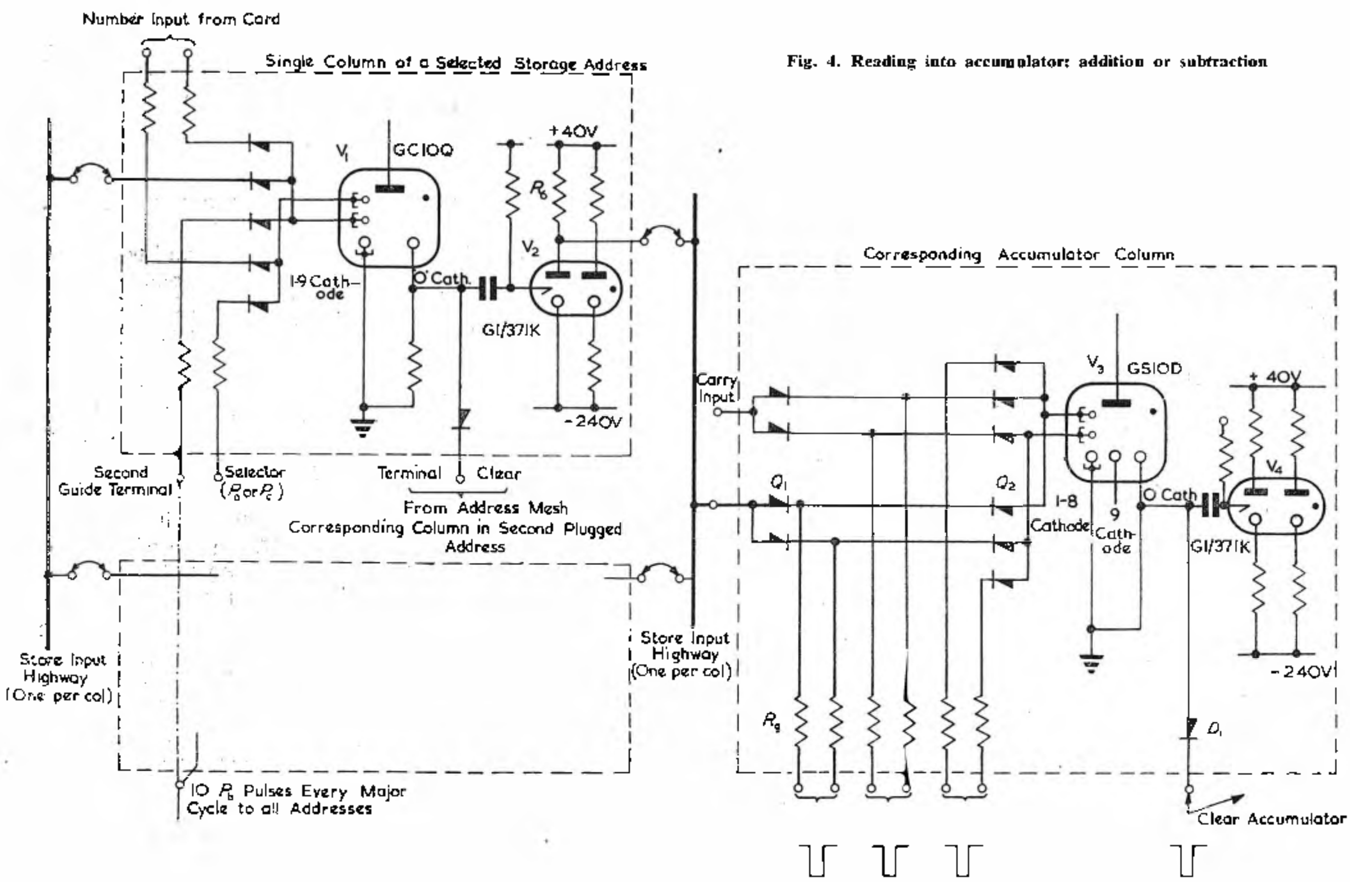


Fig. 4. Reading into accumulator: addition or subtraction

The nine complement of x has been added to the accumulator. Again, as in the case of addition, if $(9 - x) + y \geqslant 10$, a carry tube will be fired.

CARRYING AND NINE-SENSING

In the two preceding paragraphs, emphasis has been given to the production of the primary sum by a major cycle.

From the previous example, it is clear that to obtain the final sum, the primary and any secondary carries must be added into the appropriate columns of the accumulator, in subsequent carry cycles.

Every time there are ensuing secondary carries from one carry cycle, produced by the presence of a nine in a column of the sum into which a carry is added, another carry cycle is required for their addition; the worst case being that with all the columns but the lowest, being nine, and a carry being added to the second lowest column. This would require the maximum number of carry cycles.

There would be difficulties, also, with the extinguishing of the trigger tubes between each of these carry cycles. To avoid these difficulties, it was decided to use instead a system of nine-sensing analogous to that used on many mechanical accounting machines. Stated simply, this system senses the presence of a nine in any column (n), and if there is a carry from a preceding column ($n - 1$), a 'one' is added, not only to column (n) but to column ($n + 1$) as well.

This can be carried further by stating that the carry signal from the preceding column ($n - 1$) can either result from a stored carry in ($n - 1$), or from a similar nine being present in ($n - 1$) with a carry from the ($n - 2$) column, etc.

The logic for this system evolves from the four possible states of an accumulator Dekatron after a number x and a number y have been added by a major cycle.

THESE ARE:	RESULTING IN:
(a) $10<x+y<19$	Trigger tube on '0' cathode struck
(b) $x+y=10$	Trigger tube on '0' cathode struck
(c) $x+y=9$	Trigger tube not struck
(d) $x+y<9$	Trigger tube not struck

Cases (a) and (b) are essentially the same, and the possibilities reduce to three,

(1) $x + y = 9$

(2) $x + y < 9$ but with trigger tube struck

(3) $x + y < 9$ without trigger tube struck.

Considering the outputs required, when a 'carry' or 'no carry' signal is applied to a column n, one obtains:

INPUT TO COL n FROM $n - 1$	STATE OF COLUMN (n)	OUTPUT OF COL n TO $n + 1$
Carry Signal in	'0' tube not struck '0' tube struck Glow on 9	No carry Carry Carry
No Carry Signal in	'0' tube not struck '0' tube struck Glow on 9	No carry Carry No carry

From an examination of this it is seen that there is a carry output,

(a) whenever an '0' tube is struck (regardless of input) or (b) whenever there is a carry input and the glow is on nine.

In logical symbols the carry circuit of one column is shown inside the dotted lines of Fig. 5(a).

It is clear that the carry output of one column is the carry input of the next, and that they will be in the form of d.c. gating potentials. Only one carry cycle is required, a 'one' being added simultaneously into any Dekatron whose carry gate is opened by a carry input. All that is required is a further pair of overlapping driving pulses (similar to P_a and P_b) generated (after a short delay to allow the logical circuits to reach equilibrium), at the end of each major cycle and fed to two separate carry pulse lines. These two are distributed simultaneously to all carry terminals of the accumulator, for gating in as required.

In Fig. 5(b) a diagram is shown of the carry circuit required for one column[3]. A diode gate per guide, quite separate from the input gates, is used to gate the carry pulse. The carry pulses are shown on the diagram, as being initiated, with a fixed delay d following the back edge of the major cycle gating waveform. This delay allows the d.c. potentials to all the gates to reach equilibrium. The carry signal comes in at W, and as in the diode gate mentioned previously, the Dekatron will only respond to the carry pulses if the potential of W is dropped from +40V to −60V. If W is at +40V (no carry signal in) valve V_{5b} (ECC81) is turned 'on' and its anode potential falls. Resistors R_6 and R_7 are chosen so that the potential at the junction is well below the 'cut off' voltage of the suppressor grid of valve V_4 (6F33).

If W drops to −60V (carry input), the suppressor grid of valve V_4 is made positive with respect to its cathode, by valve V_{5b} being 'cut off'. Valve V_4 is still 'cut off' by the bias on its control grid from resistors R_2, R_4 and R_5. If the glow in column n is standing on cathode '9', this bias is removed and valve V_4 turns 'on'. The anode potential falls and the potential at the junction of resistors R_{12} and R_{13}, which was caught at +40V by the diode D_1, now falls until caught by diode D_2 at −60V. This was the requirement of the AND gate in the logical diagram, i.e. the voltage Y drops from +40V to −60V only if there is a carry input and the glow is on '9'. Valve V_{5a} is a cathode-follower to obtain a low impedance input to the following diode OR gate.

The voltage x is controlled only by the condition of the '0' trigger tube V_2. If $x + y \geqslant 10$, the tube will have been struck and the voltage x will have fallen from +40V to −60V.

Unistors Q_3 and Q_4 form an OR gate such that voltage Z from the cathode-follower V_{3b} drops from +40V to −60V, following either voltage x or voltage y. This voltage Z is the input W of the next stage of the accumulator. The capacitor C_1 provides a degree of integration on the output of V_4 to hold the voltage S over a carry cycle, otherwise the leading edge of the first carry pulse will remove the glow from cathode 9 and the AND gate will close immediately, cutting off all carry pulses into the following stages.

Transfer of Numbers from the Accumulator to the Store

Fig. 6 shows the circuit arrangement for 'read out.'

The roles of accumulator and store Dekatrons are reversed, if one refers to the transfer described for 'read in'. The G1/371k trigger tube in each accumulator column has two roles. It acts as a temporary store for the carries, but it is also used as a gate control when reading numbers out of the accumulator. For reading out, a major cycle of P_a and P_b pulses is routed to the accumulator read out terminals. As there is no intervening gate, the accumulator Dekatrons respond to all ten parts of these pulses and the trigger tube in the accumulator column n will fire after 10-z pulses, where z is the digit stored in the column n.

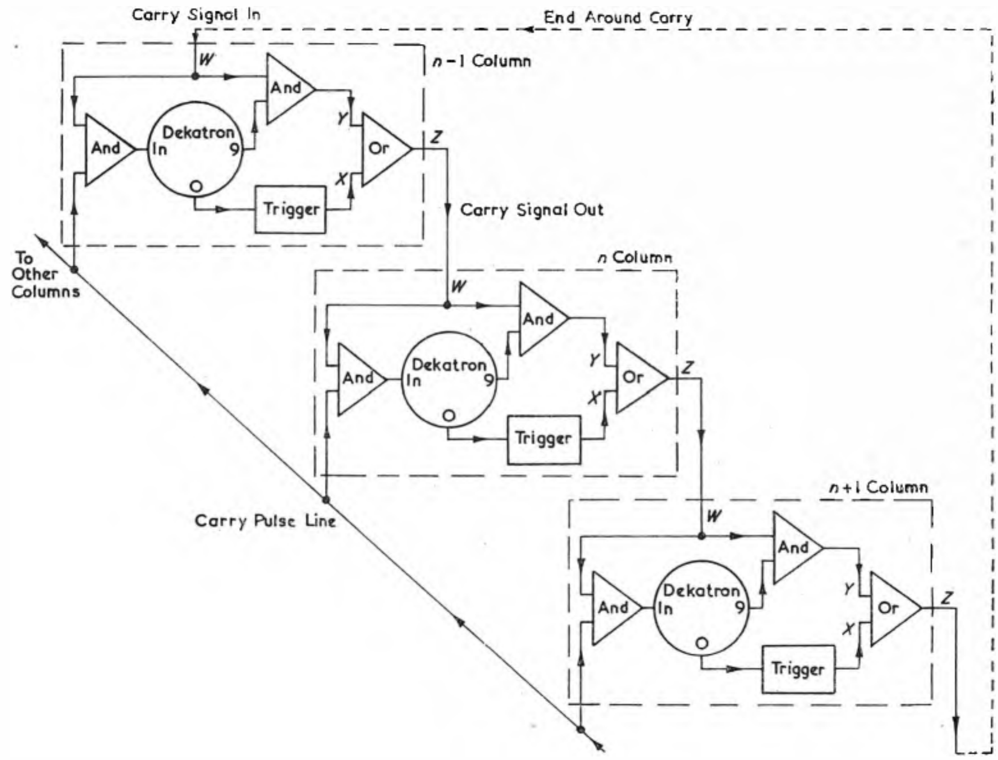
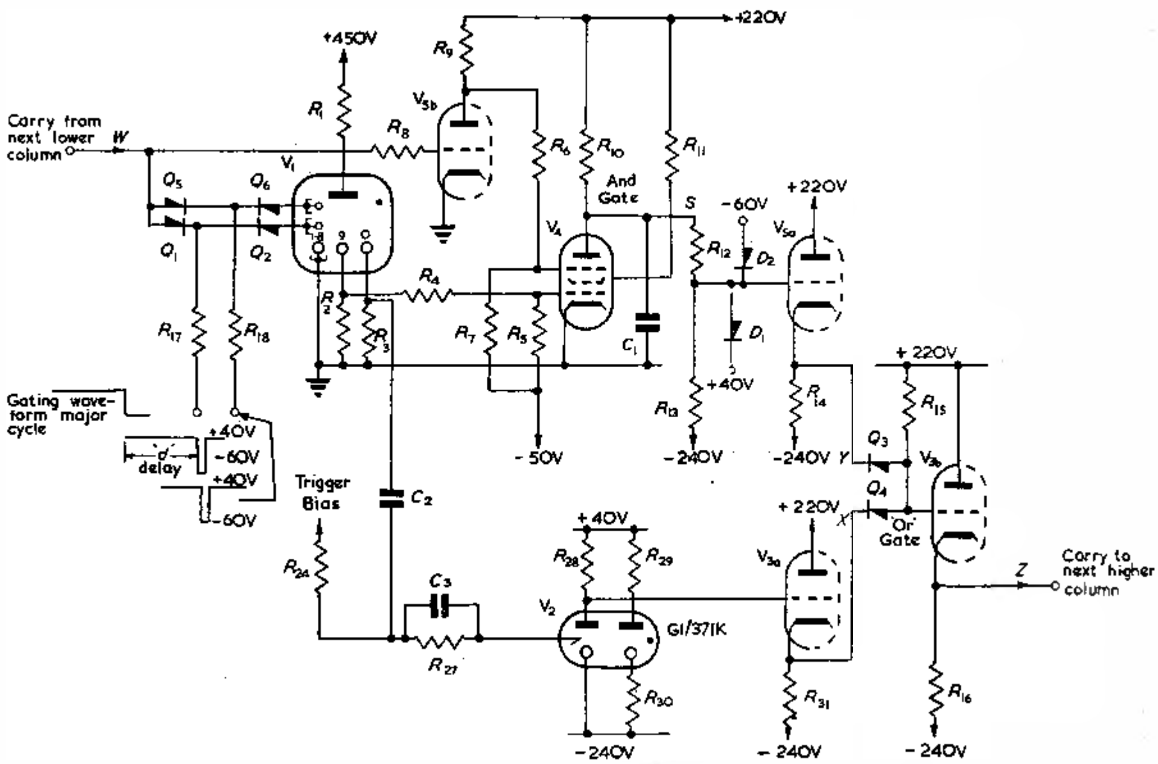


Fig. 5(a). Logical diagram of nine-sensing 'carry' system

Fig. 5(b). One column of 'carry' circuit



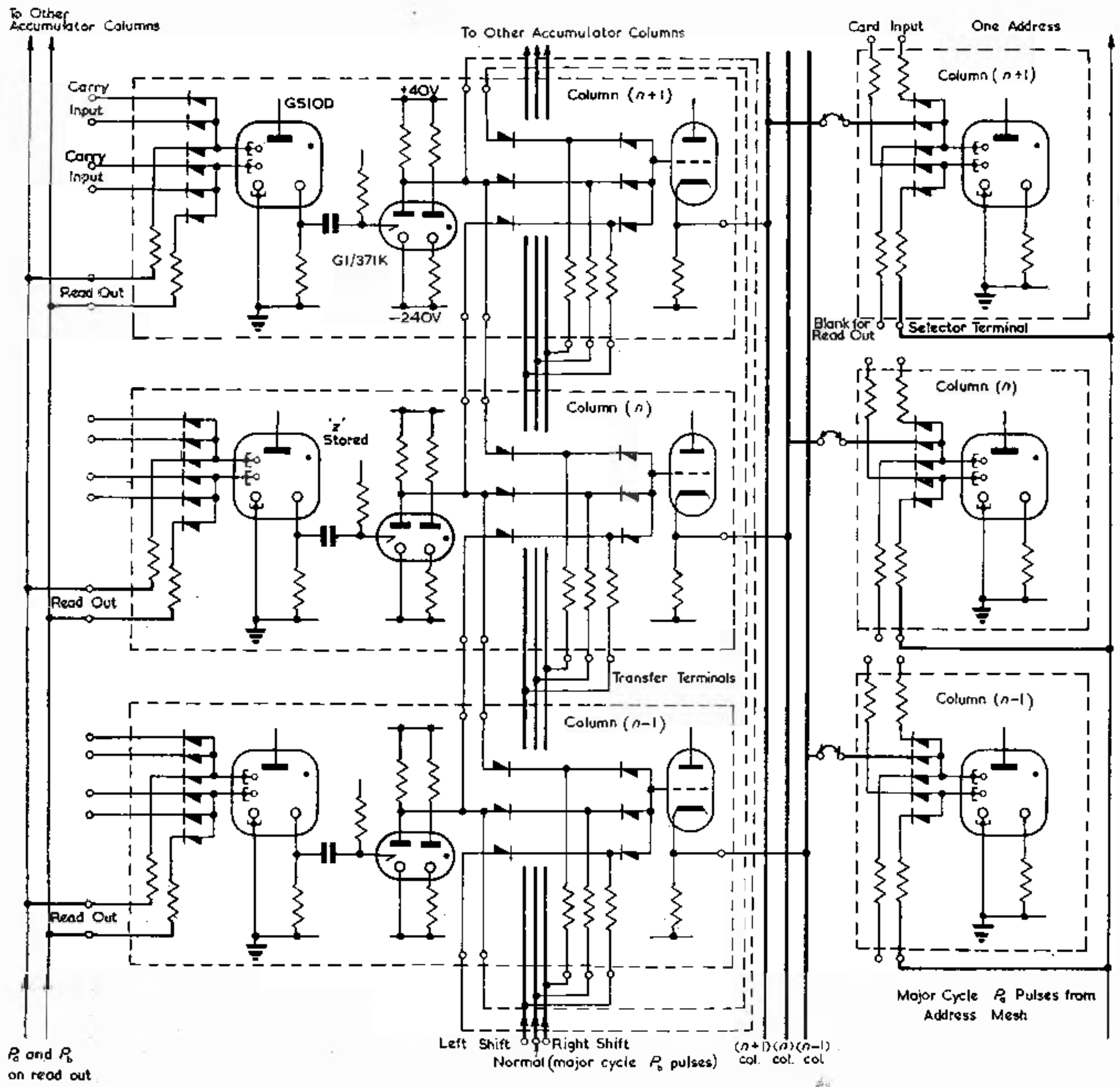


Fig. 6. Reading-out of accumulator into store

From Fig. 6 it is seen that there are three diode gates feeding the output cathode-follower in the column (n). One is controlled by the $n-1$ column trigger tube, one by the $n+1$ trigger tube and one by its own (n) trigger tube. These tubes are used to gate portions of the major cycle as before. The type of transfer is selected by the order mesh routing a major cycle of P_b pulses to the appropriate transfer terminal of the three. Gated pulses from this cycle reach the second guides of all the storage address Dekatrons plugged to that column highway via the output cathode-follower. Address selection is made by the address mesh routing a major cycle of pulses (P_a pulses always for read out) to the selected address selector terminal. The number will always be read into the store as a true number.

Normal Transfer

The P_b pulses are fed to the centre diode gate which is controlled by the trigger in the same column. This results in a column for column transfer of the number from accumulator to storage address.

Left Shift (Multiply by Ten)

The pulses are fed to the lower gates of each column. As the only lower gate that the n trigger tube is controlling is in the $n+1$ column, the digit z is fed to the $n+1$ highway. This results in multiplying the number z by ten. As this occurs for all the columns in the accumulator, the whole number in the accumulator can be multiplied by ten on reading out.

Right Shift (Divide by Ten)

Similarly, by routing the P_b pulses to the upper gate in every column, the number z is fed out into the $n-1$ highway, because the only upper gate that the n trigger tube

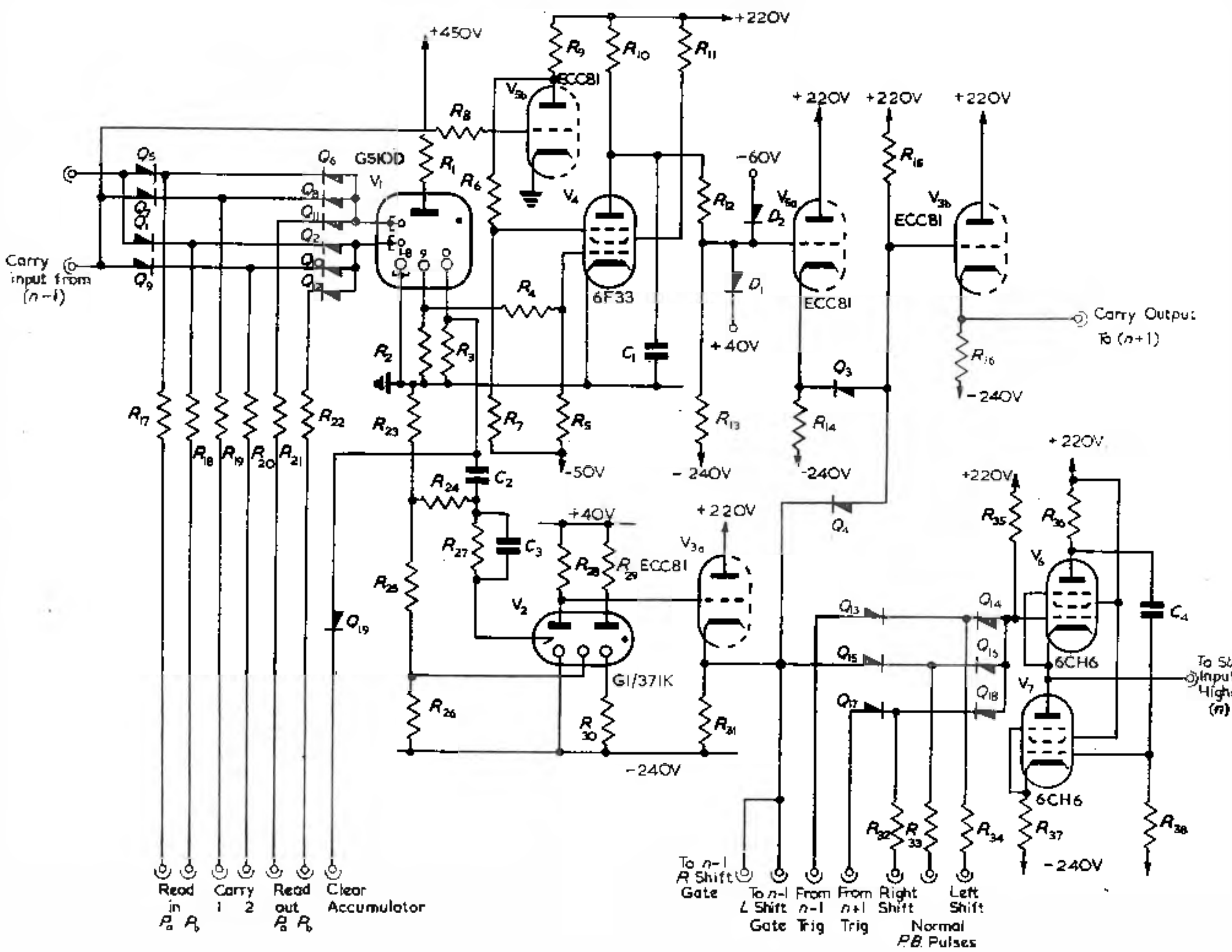
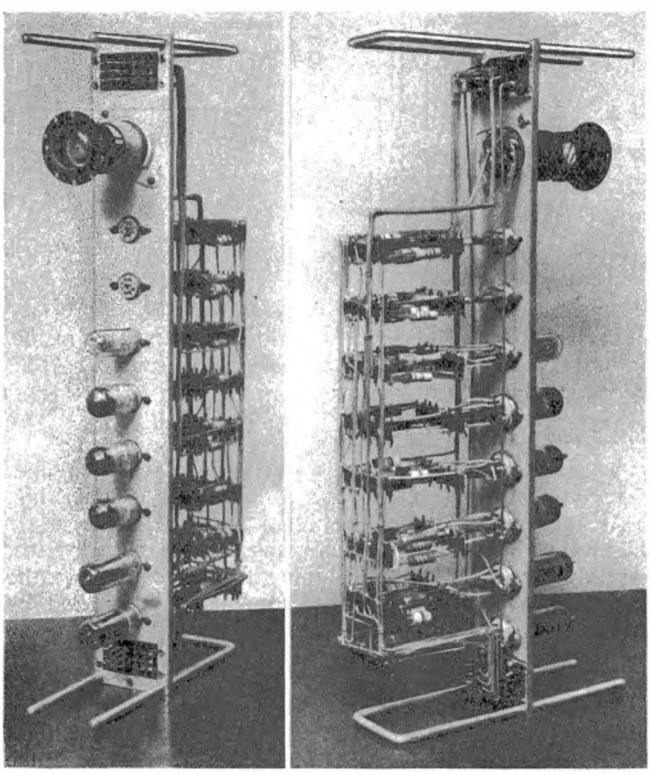


Fig. 7. Circuit of one column (n) of accumulator

Fig. 8. One column stage of the accumulator



is controlling, is in the $n-1$ column. The whole number can thus be divided by ten on reading out.

Construction

Fig. 7 shows the complete circuit of one column chassis of a Dekatron accumulator. These are standard plug-in units, making an accumulator of variable length. A turret construction is used for assembling the circuit components around the valve base. The units are plugged in vertically so that the Dekatrons form a horizontal line for easy viewing. Two aspects of the chassis are shown in Fig. 8.

Conclusions

At the expense of five hard valves, one cold-cathode trigger tube and a Dekatron per stage, this accumulator can add, subtract, multiply or divide by ten in a basic time of 550μsec. The accumulator has the further advantage that it can be read directly at any time, which assists the service engineer and operator alike.

Acknowledgments

The authors wish to express their thanks to the Directors of the British Tabulating Machine Company Limited for permission to publish this article.

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Heat Control in Electronic Equipment

(Part 2)

By E. N. Shaw*, A.M.I.E.E.

Introducing Conduction

The exploratory measurements having demonstrated the possibility of controlling the heat flow from a 'floating' heat source into a selected surface, it seemed logical to augment this flow by providing a conductive path to the surface capable of the most effective dissipation by all three methods of propagation. Conveying heat by conduction to the rack with which it is in contact, the front panel is an obvious choice with the promise of high convection loss from a large vertical surface that could be given a high-emissivity finish to increase radiation loss. Mechanical considerations require all components to be supported from the front panel, with the three other surfaces constituting a dust cover to which nothing may be attached, so this choice in no way clashes with conventional methods of construction used for several decades.

A conductive path was therefore provided between the heat source and the front panel in the usual way—but not necessarily for the usual reason of purely mechanical necessity.

Conventional Construction

Plate 5 shows the results of experiments on several units with a conventional 'inverted tray' chassis extending the entire length of the enclosure and in light contact with all vertical surfaces of the container except the front panel, to which it was firmly bolted to ensure good thermal contact. The heat source was mounted on the upper surface of the chassis with the valve bases and resistors passing through an aperture to appear, traditionally, in the space below. Great care was taken to make good thermal contact between the valve deck and the chassis in order to offer low impedance to heat flow by conduction into the front panel with the chassis kept clean to minimize radiation loss to the interior of the container. Mechanical arrangements concerned with the mobile thermo-couples restricted the choice of height for the chassis, but a position was found such that the valves were approximately half way between top and bottom of the container enabling comparisons to be made with previous experiments with the source in an intermediate position.

Comparing Exp. 28 with Exp. 20 brings to light several interesting differences. Care taken to achieve good heat flow by conduction into the front panel was rewarded by a temperature drop of no less than 71° for the deck, together with 15° for the valves, 5° for skin average and 4° for rack average. The isothermal pattern above the chassis dropped by some 20°, while the ambient in the vicinity of the load resistors remained about the same because the pattern assumed the stratified form associated with heat-dissipating components immediately below a horizontal surface with no possibility of relief from stagnation by the formation of even a frustrated convection current (see Exp. 19 and Exp. 22).

It was expected that reductions in container temperatures would have to be paid for by a hotter rack, but the 4° reduction proved the opposite to be true. The explanation lies in the fact that, while average skin temperature dropped by 5°, the expected rise appeared at the junction point of chassis and front panel, which grew hotter by 13°. Supplied via a low-impedance path from the heat source, the temperature recorded at this level suffered very little attenuation along the length of the panel and provided a larger area of hotter surface from which heat was removed at a greater rate by convection, leaving less to flow into the rack.

Painting black all surfaces (except the chassis) confirmed the exploratory measurements by dropping the isothermal pattern by 40° without appreciable change in shape. Comparing Exp. 29 with Exp. 23 shows a temperature drop of 62° for the deck, together with 14° for the valves, 4° for skin average and 3·5° for rack average, with only 9° rise at the junction of chassis and panel because radiation aided dissipation by convection. It should be noted that introducing conduction within a container with high-emissivity surfaces reduced the deck temperature rise to that recorded in free air.

Moving on to the all-black-and-ventilated container of Exp. 30, one may be pardoned a moment of surprise at finding the isothermal pattern similar to those observed in totally enclosed containers. No appreciable impedance to the escape of hot air was offered by the perforated metal top, yet the pattern above the chassis was no less frustrated, with the stratified condition maintained below the chassis despite low impedance to the entry of cold air through the perforated metal base.

This is, of course, a classical example of a container designed for 'straight through' ventilation being made virtually useless by the baffling action of the vast expanse of horizontal surface associated with an 'inverted tray' chassis. Exp. 30 being a little cooler than the previous one suggests that there must have been some ingress of cold air at points remote from the hotter central section, although there can be little doubt that it was mostly 'lazy' air, as is shown by comparison with Exp. 26 in which the air was anything but 'lazy'. Introducing conduction in this instance reduced the temperature at one point only—the valve deck, which was 13° cooler than in Exp. 26. Restricting the ventilation raised valve temperatures by 12° and forced the rack average up by 10° in an effort to lose more heat by radiation and convection from the external surfaces to which heat had been conducted.

No apology is offered for including among the experiments such an illogical arrangement. It is, regrettably, representative of standard practice. It might even be considered as being slightly advanced, in that 'straight through' ventilation could have been achieved by suitable chassis design.

Punching random holes in the chassis is not the answer because they would encourage 'lazy' air without reducing stagnation below the chassis, the only effective way of clearing which is to provide apertures directly above every heat-dissipating component. Convection streams issuing from such apertures would be so intent upon escaping upwards that the valves would derive no benefit from them and would require individual supplies of cooling air from a circle of holes round each valve, which would raise the impedance of the conducting path to the panel. Compromise with an 'inverted tray' chassis will clearly not produce optimum conditions, which will obtain only when

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full and simultaneous use is made of both means of heat transfer by departing from conventional practice.

Reverting to conduction alone, the considerable temperature reductions recorded in Exps. 28 to 30 could have been improved upon by moving the valves nearer the panel to reduce the length of conducting path and loss along it by radiation to the interior. Such a move would have also increased radiation transfer to the panel and reduced that directed at the rear of the dust cover. Nothing could have been done, however, about the heat flow by conduction along the expanse of chassis between the heat source and the rear of the dust cover. This flow is not only opposite to the desired flow into the front panel, but is also completely aimless, having to be dissipated by radiation into the interior of the container. From this it follows that the ideal internal structure will have to eliminate 'aimless' conduction transfer.

Conventional Units in a Rack Assembly

The lower portion of Plate 5 indicates what happened when the three forms of conventional unit discussed above were mounted on a rack with other units of the same type. Detailed measurements were made on the central unit, fully loaded to the standard dissipation of 178W, with a 'hot' unit below identical in finish and dissipation. The upper unit was left unloaded, it being desired to show that a passive, or 'cold' unit had an appreciable effect on the heating of a unit below it. This series of experiments was devised to show the means by which heat transfer occurred between units rather than to record the resulting temperature rise, which would have been much greater with fully-loaded upper units.

Compared with the singly-mounted Exp. 28, the central unit of Exp. 31 is seen to have an isothermal pattern similar in form, but elevated by some 20°. Cooling was mainly by convection from low-emissivity vertical surfaces and conduction to the rack, the horizontal surfaces being denied contact with free air. As a result the average skin temperature rose by 17°, with the greatest rise (29°) at the lower surface, which was not only denied contact with free air, but had under it the hot upper surface of the lower unit which, however poor its emissivity, was sufficiently hot to produce this rise by radiation. Warming of the 'cold' unit took place by a combination of radiation from the upper surface of the central unit and conduction via the rack, which was itself warmer by an average of 12°.

The high emissivity of the black-painted units in Exp. 32 had the expected effect of reducing the isothermal levels by 40°, although these were still about 20° higher than in the singly-mounted Exp. 29. Relying upon cooling by convection aided by radiation from high-emissivity surfaces, it was expected that denying this facility to the upper and lower horizontal surfaces would have a greater effect than when dealing with convection loss alone. This proved correct, the average rise in skin temperature being 19°, with a maximum of 37·5° at the bottom of the container where the high-absorptivity surface (no longer emitting) accepted heat from the high-emissivity top surface of the unit below. This was an involuntary betrayal of function by a surface which, intended to be a good radiator, was prevented from performing this function and forced into accepting heat from an equally good—but appreciably hotter—radiator in close proximity to it. It speaks volumes for the high efficiency of the unrestricted vertical surfaces that they were able to cope with this influx of heat without calling upon the rack to heat up by more than 15·5°.

The radiating property of the upper surface of the container was not entirely lost. Denied the opportunity of radiating into space, it grew hotter as a result of heat reflected from the clean surface of the 'cold' unit, which absorbed sufficient energy to raise its temperature by 22°, some of which may have been due to contact with the rack. Painting black the underside of the 'cold' unit to increase its absorptivity raised its temperature by 6° and dropped that of the upper surface of the central container by the same amount, indicating that the two surfaces were acting as more efficient 'heat exchangers'. Complete removal of the 'cold' unit cooled the upper surface of the container by 18° and the rack by 3°, with proportional reductions throughout the whole container, these minor variations demonstrating the fact that conditions within a totally enclosed container are influenced by the presence, and surface finish, of a completely passive unit mounted above it.

Comparison of temperatures in singly-mounted units with those in the same units sandwiched shows similar increases which, taken in conjunction with the general rise in isothermal levels by some 20° in each case, suggests that there is little to choose between the two. Actually, if these differences are related to the dissimilar starting levels in the singly-mounted units, it is seen that, for example, deck and rack temperatures increased by 8·7 per cent and 31 per cent respectively in Exp. 31—and by 12·8 per cent and 63 per cent in Exp. 32.

On this basis of comparison the effect of sandwiching a high-emissivity container is shown to be appreciably greater.

Applying the same method of comparison to Exp. 33 reveals a rise of 38 per cent in deck temperature and 105 per cent in rack temperature. It is not difficult to attribute this sharp rise to the unfortunate combination of ventilated container and baffling chassis. The stagnant conditions below the chassis that prevented the entry of cold air in Exp. 30 were disrupted in Exp. 33 by hot air forcing its way in from the unit below. Passing through the two surfaces of perforated metal, the hot air was frustrated by the baffling action of the chassis, to which it gave up some of its heat before escaping through the inter-unit space. The frustration level was thus transferred from the top of the lower unit to the container chassis, raising the ambient below it by 40° and increasing deck and valve temperatures by 31° and 15° respectively. Isothermal levels above the chassis also rose by 40°, with the sluggish convection stream passing through the perforated metal top to be frustrated by the solid black base of the 'cold' unit, to which it imparted enough energy to raise its temperature by no less than 39°. Changing to a clean finish on the base of the 'cold' unit reduced this rise by 5° and raised the isothermal level within the container by the same amount, while complete removal of the 'cold' unit eased conditions in the upper part of the container by 20°.

If, as was suggested earlier, 'straight through' ventilation had been provided by punching random holes in the chassis, the greater impetus of the convection stream would have transferred more heat from the lower unit into the central container and the frustrating 'cold' unit would have suffered an even bigger temperature rise.

Factors Affecting Heat Transfer Between Units in a Rack

From these experiments on conventional units two contradictory conclusions can be drawn. The first is that very appreciable temperature reductions can be achieved in singly-mounted units by introducing a conductive path to the front panel, increasing surface emissivity and providing a modicum of ventilation. The second is that these progressive improvements are nullified when the units are sandwiched between others in a rack, each 'improvement' resulting in greater susceptibility to heat transfer from the unit below and increasing ability to convey 'unwanted' heat to the unit above. As an example of the 'swings-and-

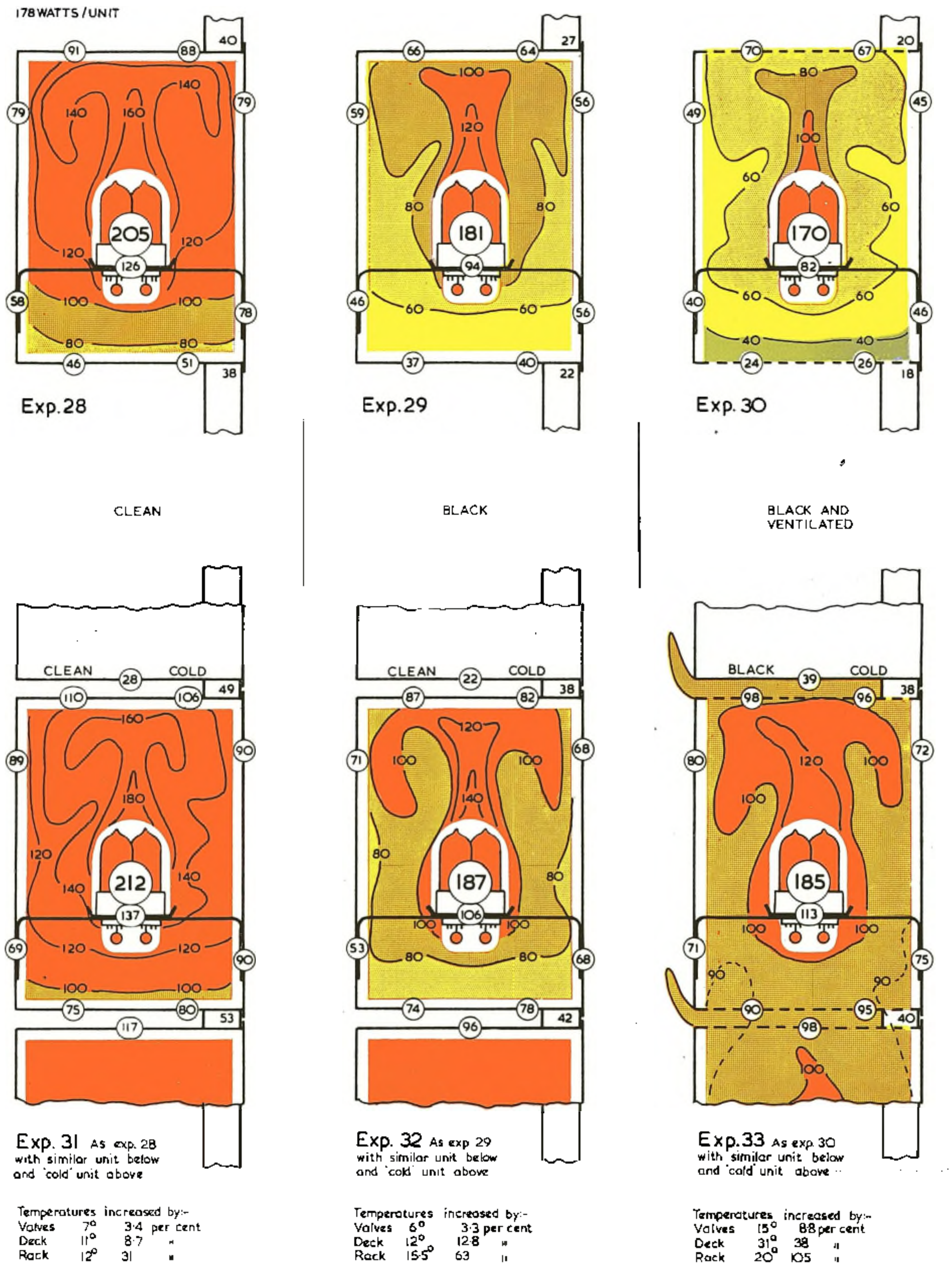
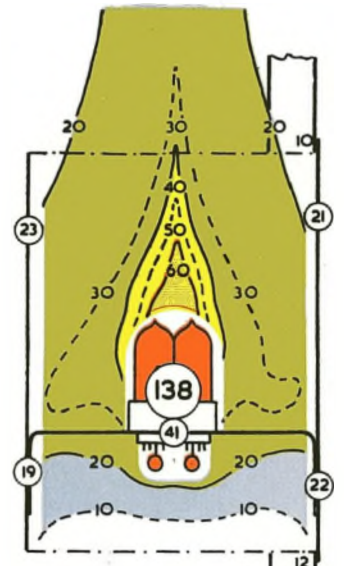
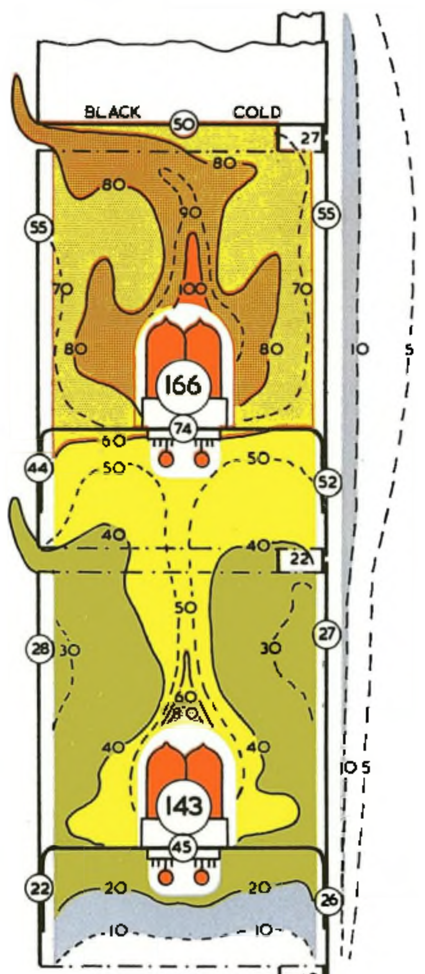


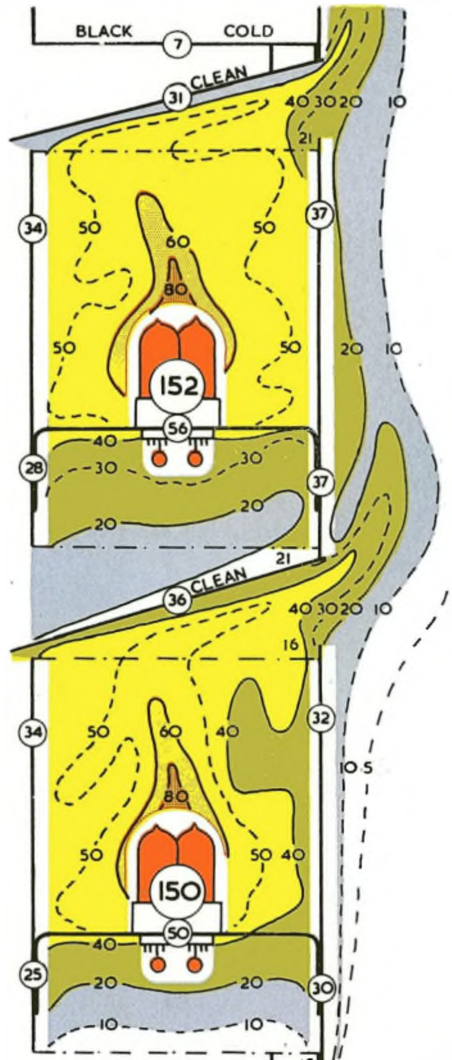
Plate 5. Conventional units mounted on a rack



Exp. 34. Special unit in an all-black container without upper or lower horizontal surfaces.
Layout of heat source as in Fig. 5, but 'straight through' ventilation provided by ten $\frac{1}{2}$in dia. holes in the chassis, one directly above each of the load resistors. Total dissipation 178 watts

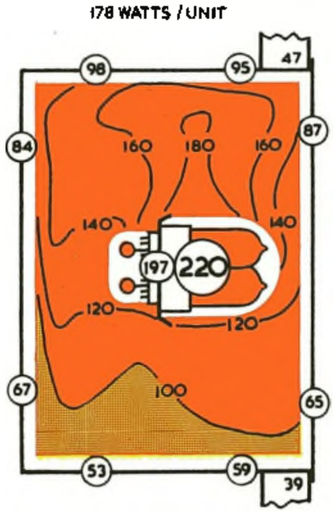


Exp. 35 Example of 'chimney effect' when well-ventilated units are mounted one above the other. Introducing deflectors provides an interim solution →

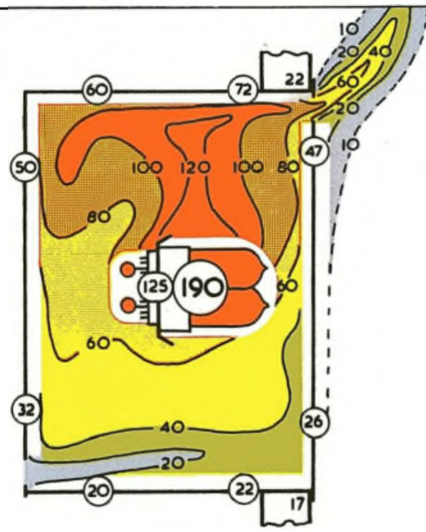


Exp. 36 Deflectors in $1\frac{3}{4}$in inter-unit spaces

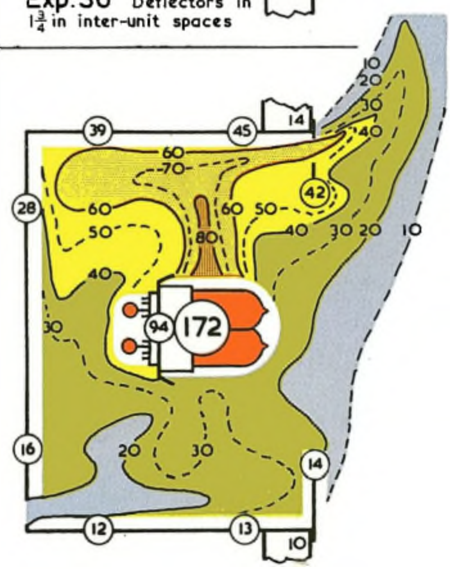
178 WATTS / UNIT



Exp. 37 As Exp. 20, inner and outer surfaces clean



Exp. 38 $\frac{1}{2}$in slot 11in long in front panel and dust cover



Exp. 39 Aperture 5in high and 11in long in front panel

Plate 6. Interim solution and start of final design

roundabouts' principle besetting all heat problems, this dilemma doubtless accounts for the lack of progress in the design of rack-mounting units so long as they are subjected to unpredictable thermal hazards.

A logical approach to the way out of this dilemma is to examine the mechanism of inter-unit heat transfer under its three main headings, viz:

Conduction

The front panel of each unit is in thermal contact with the rack and conducts heat to it. Ideally, the contribution from identical units will be uniform and there should be no heat transfer between units by conduction in the absence of a temperature gradient. Experience and experiment show that this ideal is not attained, there being a progressive increase in unit temperature with height accompanied by a corresponding build-up of temperature gradient along the rack. Whatever the cause of the upward trend, it is definitely not conduction, which can only counteract the general tendency by heat flow 'down gradient' in an attempt to restore equilibrium. In addition to this equalizing function the rack dissipates heat by radiation and convection, acting as a heat sink for hot units and as an unavoidable heat source for 'cold' units. On balance, conduction by the rack is more beneficial than otherwise, which is just as well when it is realized that nothing can be done to control it.

Eliminating conduction as a primary means of heat transfer between units does not mean that it is valueless within containers. On the contrary, if inter-unit transfer by other means is to be minimized, the general temperature level within each container must be reduced by all possible expedients, of which conduction is one of the most effective. Moreover, heat flow by conduction is under the direct control of the designer, whose aim it must be to provide a low-impedance path between the heat source and an external 'heat sink' in the form of a surface capable of dissipating heat in the required direction. The front panel is such a surface and the rack (failing conduction along it) may be considered to be an extension of the front panel with the combination forming a 'heat sink' dissipating heat by radiation and convection.

Comparing Exps. 28, 29 and 30 with their earlier counterparts showed marked temperature reductions due to the introduction of a conducting path between the heat source and the front panel, with the possibility of even greater reductions if steps had been taken to eliminate 'aimless' conduction in the opposite direction and to reduce the path impedance by moving the source nearer the panel.

Following this process to its logical conclusion suggests that optimum conditions would obtain with the conventional chassis removed and the heat source mounted directly on the front panel.

Radiation

Several experiments have shown the value of high-emissivity surfaces as aids to heat dissipation by radiation, improvements of at least 40° in isothermal levels being obtained within singly-mounted units in which all surfaces were free to radiate. It is unfortunate that sandwiching such units between others in a rack not only inhibits radiation loss from all horizontal surfaces, but encourages the most unwelcome upward heat flow by providing efficient 'heat-exchangers' between units, with the high absorptivity lower surface of each container accepting heat from the high-emissivity radiating surface of the hot unit below.

The amount of heat exchanged between two such surfaces depends on their separation, emissivity and temperature differential, of which the first is fixed by long-established standards for the dimensions of dust covers in relation to front panels. Both other factors are, however, under the control of the designer, whose first step should be to reduce the emissivity of the surfaces in juxtaposition.

This could have been demonstrated by removing the black paint from all horizontal surfaces in Exp. 32 or, alternatively, by painting black all vertical surfaces in Exp. 31. The result would have been intermediate between the two, with radiation transfer between clean surfaces reduced (as in Exp. 31) but not eliminated because zero emissivity is unattainable and the temperature differential between the two surfaces would have remained high.

It is the reduction of this temperature differential (without which heat transfer cannot occur) that is the major problem confronting the designer wishing to minimize radiation transfer in a rack assembly.

In theory, the solution to this problem is simple. It must be recognized that horizontal container surfaces cannot (and must not) be used as convecting or radiating surfaces, with heat flow so controlled that none reaches these surfaces by any of the three means of propagation. In other words, the dust cover must be just what its name implies, remaining thermally neutral and playing no part in the process of heat extraction from the interior, which must take place only via the front panel.

In practice, the stringent requirement of a thermally neutral dust cover is not easy to meet. Conduction transfer can be ruled out because the dust cover is not normally in thermal contact with the heat source and any flow from the front panel can be reduced to negligible proportions by a strip of insulating material along the line of already indifferent contact. Radiation heating can also be reduced by ensuring that internal surfaces are of poor absorptivity and by moving the source nearer the front panel.

It is the transfer of heat by internal convection that is most difficult to control because, whatever the position of the source within the container, some of its heat is caught up by the irrepressible convection current and flung against the frustrating top surface of the dust cover to produce the very temperature differential that it is desired to abolish.

Convection

Heat transfer by convection is present in three forms. Exposed surfaces of rack-mounting units are cooled by *external* convection, the amount of heat dissipated being a function of the swept area and temperature rise to the power 4/3. With a thermally neutral dust cover the front panel is the only surface capable of dissipation and it is profitable to make the best use of the practically limitless supply of cold air traversing it by making the area of the front panel large and raising its temperature by every possible means of heat transfer from the container interior. It has been shown that both conduction and radiation can be controlled to give an increased heat flow horizontally into the front panel, although heat transfer by *internal* convection remains most adamantly vertical.

This confirms a previous conclusion that *internal* convection with a limited supply of air contributes little to heat extraction from a container, although it does relieve local 'hot spot' conditions by transferring heat to the top of the dust cover. Not only is this transfer in a direction at right-angles to that desired, but it is also remarkably potent, the temperature of the upper surfaces of the dust cover being appreciably higher that that of the panel in every one of the experiments so far attempted!

Faced with the impossibility of suppressing *internal* convection, the designer is tempted to make a virtue of necessity by changing his approach to the problem and concentrating upon direct cooling of the source by increasing the vigour of convection loss from within the container. This can be done by providing vents in the upper surface of the dust cover to remove the frustration experienced by *internal* convection and admitting the limitless supplies of cold air associated with *external* convection through similar vents in the lower surface, the resulting *ventilation* combining the two forms of convection.

The exploratory measurements recorded on Plate 4 show how effective 'straight through' ventilation can be, but Plate 5 tends to destroy any optimism by demonstrating that the conventional 'inverted tray' chassis has a serious baffling effect in a singly-mounted ventilated unit (Exp. 30) and an even more serious frustrating effect (Exp. 33) in an assembly of similar units. Removal of this frustration would convert the assembly into a 'chimney' with the combined convection currents sweeping upwards bearing ever-increasing quantities of heat from progressively hotter units. As this condition is frequently met in practice it is not surprising to find low survival-rates in components and valves, especially if a carelessly placed horizontal surface frustrates this buoyant avalanche of thermal energy and transforms the assembly into a large-scale version of the stagnant and stratified Exp. 19.

Conclusions Leading to an Interim Solution

Examination of experimental results and their analysis can be summarized by the following conclusions:

(1) Inter-unit heat transfer is not aided by conduction along the rack.

(2) Radiation transfer between totally enclosed units is due to differential heating of horizontal surfaces of the dust cover by internal convection, the upward flow of heat being in the form of a series of convection currents interrupted by 'heat exchangers.'

(3) 'Straight through' ventilation is the most direct and effective method of cooling the heat source, but it is also the most direct and effective means of progressive heating of rack-mounted units by 'chimney effect.'

(4) If the undoubted advantages of 'straight through' ventilation are to be fully exploited, hot air leaving a unit must not be allowed to enter the unit above.

Giving the last of these conclusions the force of a directive determined the next step to be taken, either by experiment or by logic. Choosing the latter, Exp. 33 was re-examined to see what could be done to reduce 'chimney effect.' It was noted that hot air from the upper container escaped via the inter-unit space without entering the 'cold' unit, although frustration caused by the violent change in direction raised the temperature of the latter by 39°. It was thought that this heating could be reduced by a clean metal sheet interposed between the units that would act as the frustrating surface without impeding the escape of air. Transferring attention to the perforated base of the central container, it was seen that hot air from the lower unit was frustrated by the chassis before it escaped through the inter-unit space. It was thought that a similar interposed plate would form the frustrating surface and prevent hot air from entering the central unit, which could then be modified by piercing holes in the chassis to increase the vigour of ventilation by admitting cold air.

The fallacy in the last item of reasoning is that a single aperture cannot be used both as inlet and exhaust. As the front panels form a practically continuous surface, the hot air from inter-unit spaces can escape only towards the rear of the assembly and cannot avoid being sucked into the inlets of the units above.

This difficulty could be resolved by moving the units apart by one modular increment ($1\frac{3}{4}$in) of the rack-mounting system to provide alternative apertures in the front of the assembly to act as inlet or exhaust. Choice of function would be under the control of the designer, who could tilt the interposed metal sheets in either direction, the logical choice being to deflect the rising columns of hot air towards the front of the assembly where they would join the external convection current sweeping upwards across the hot surfaces of the front panels.

Experimental Verification

The validity of the above deductions was checked by experiments shown in the upper part of Plate 6. Not wishing to modify the existing heat source, two special units were constructed with the valves mounted directly on the chassis, using exactly the same layout as in Fig. 5 except that a $\frac{1}{2}$in dia. hole was provided immediately above each load resistor to remove local frustration and to cool the valves by 'straight through' ventilation, this assembly being sufficiently compact to avoid the necessity of individual cooling arrangements for each valve. All-black containers were used with the upper and lower horizontal surfaces completely removed to avoid offering any additional impedance to the anticipated high-velocity air flow.

Living up to expectations, the singly-mounted unit of Exp. 34 returned the lowest valve and deck temperatures recorded throughout the entire series of experiments, with a smooth isothermal pattern suggesting that the ventilating holes had been correctly placed for cooling the valves as well as the load resistors. This was checked by placing screening cans on the valves, which immediately rose by 24° to 162°, with the increased conduction flow through the chassis raising its temperature by 25° to 66° above room ambient.

Removing the screening cans and placing a similar unit below and a 'cold' unit above reproduced, in Exp. 35, the 'chimney effect' so often found (and sometimes even encouraged) in assemblies of rack-mounting units. With its unimpeded supply of cold air, the lower unit suffered least, disgorging heated air into the under part of the sandwiched unit where frustration took place despite more than adequate chassis ventilation, forcing some of it to escape via the inter-unit space. Denied a supply of cold air, the sandwiched unit had to cope with the unwanted heat from below as well as trying to rid itself of heat by internal convection past the frustrating surface of the 'cold' unit above. Valve, deck and rack temperatures rose by 20, 80 and 123 per cent respectively when compared with a singly-mounted unit—a result inferior to preceding experiments on sandwiched units, although the actual recorded temperatures were lower than before. The price paid for this seeming improvement was the rise in base temperature of the 'cold' unit to a new maximum of 50°, compared with the 39° attained in Exp. 33, substantiating the premise that 'chimney effect' would transfer heat upwards with maximum efficiency, not unmixed with the possibility of devastating increases in temperature when frustrated.

Faced with an alarming rise in temperature of the 'cold' unit due to 'chimney effect', steps were taken to reduce it by implementing the deductions made earlier. All three units were separated by the $1\frac{3}{4}$in modular increment and inclined metal sheets placed in the inter-unit spaces to act as deflectors. If the previous experiment gave an impression of somewhat uneasy balance between credit and debit, Exp. 36 succeeded in eliminating 'chimney effect' and producing the following unqualified improvements:

(1) The temperature rise of the 'cold' unit was reduced from an alarming 50° to a bearable 7°.

(2) The sandwiched unit grew cooler by an average of 17°.

(3) The maximum difference between the two hot units fell from 29° to 7°.

That these marked improvements were due to the deflector plates rather than to increased separation was shown by removing the upper deflector plate, whereupon the 'cold' unit rose to 26°, with a further rise to 36° on removal of the lower deflector.

Advantages of the Interim Solution

Outstanding among the advantages to be derived from the interim solution is the possibility of applying it to perfectly standard units of conventional design. Any container located at the top of an equipment rack showing signs of distress can be protected from the upward heat flow by placing a deflector plate below it, while a heavily loaded unit at the base can be prevented from heating an entire assembly by a deflector plate above it. Maximum protection would, of course, be obtained by interposing deflector plates between all units to ensure that each has an individual supply of cold air which, when heated, cannot enter the unit above.

The benefits each unit derives from its individual supply of cooling air would naturally depend upon provisions made for 'straight through' ventilation, so that the user of conventional equipment in a rack assembly can not only protect units without modification, but can also increase the cooling of a protected unit to an extent determined by his willingness and ability to perform comparatively minor modifications in the certain knowledge that improving ventilation will not increase 'chimney effect', as would be the case without deflector plates.

Disadvantages of the Interim Solution

With space at a premium, the addition of a minimum of 1¾in to each unit is not likely to be welcomed, although it may tip the scales by comparison with the bulk, cost and complexity of most systems of forced ventilation. The reason for regarding 1¾in as a minimum can be understood by examining Exp. 36 in which the acute angle formed by the deflector plates does not favour the smooth exit of ventilating air, as is evident by the amount of frustration due to violent changes in the direction of air flow, nor does the necessarily clean surface prevent radiated heat from the source being reflected back into the container. It would undoubtedly be beneficial to make the deflector angle more obtuse, say 45°, so as to reflect all the radiation into outer space and decrease frustration by less violent changes in direction of air flow. To attain an angle of 45° in Exp. 36 would have needed an inter-unit separation of 6in—a very high price to pay for an approach to the conditions in Exp. 34, in themselves by no means ideal.

Approach to the Final Design

Sufficient experience had been gained at this juncture to permit the formulation of requirements to be met by the final design:

(1) The unit must be self-cooling when used on a bench.

(2) The method of cooling must not permit heat transfer by 'heat exchanger' or by 'chimney effect' when a unit is sandwiched between similar units on a rack.

(3) All undesired effects must be eliminated without any departure from the usual shape or dimensions associated with rack-mounting units and without resorting to interposed deflectors or increased spacing.

To meet these very stringent requirements the final unit will have to incorporate a 'built in' ventilating system with an internal layout that will extract heat from the interior by the most effective use of all three means of propagation without allowing these same means to transfer heat to the dust cover—a task that requires the maximum use to be made of every lesson learned in the preceding thirty-six experiments.

Evolution of the Final Design

Whereas previous experiments were devised to highlight different aspects of heat dissipation by comparison and contrast, this last phase was planned to form a logical sequence of events in the evolution of a final design. With the concept of a thermally neutral dust cover fresh in mind, a start was made by selecting from the exploratory measurements group a unit having a clean finish on all surfaces, with the firm resolve to keep the upper and lower surfaces inviolate until the end. The choice fell naturally on Exp. 20, in which the heat source (by a process of clairvoyance which need not be dwelt upon) happened to be in the 'standard' position which was kept equally inviolate throughout this final phase.

Re-drawn for easy reference as Exp. 37 in the lower part of Plate 6, the diagram shows a dust cover (and front panel) potentially neutral, but actually extremely hot because the enclosure formed a complete barrier between heat source and outer air. As the heat source was mounted in such a manner that heat loss by conduction could not take place, the surfaces of the container were heated by radiation and internal convection, with the latter responsible for the 40° differential between upper and lower surfaces of the dust cover.

Prohibited from relieving the furnace-like conditions by the use of 'straight through' ventilation, a compromise was made by cutting an exhaust vent in the form of a slot ½in wide and 11in long at the highest point on the front panel, with a similar inlet vent as low as possible in the rear surface of the dust cover, the intention being to simulate the controlling action of inclined deflectors by forcing heat-laden air from the interior to join the external convection stream without the possibility of it entering a unit above. A battery of mobile thermo-couples forward of the front panel recorded the success of this move in Exp. 38, in which the upper vent belched out a torrent of heated air at considerable velocity, as indicated by its departure from the vertical, with a 60° drop in internal isothermal levels as a measure of the quantity of heat removed. A point of interest is that the internal pattern, instead of adopting a diagonal trend from inlet to outlet, remained obstinately similar to that observed earlier, with the anti-clockwise internal circulation forcing the indrawn air to remain close to the lower surface of the dust cover until it could rise between the heat source and the front panel. The considerable temperature reductions achieved do not permit this to be called 'lazy' air, although the path it was forced to take resulted in a greater differential and left the front panel cooler than the rear of the dust cover—both effects being the opposite of those desired.

This possibility of moving a body of cold air horizontally could not be exploited without preliminary modifications involving the seemingly illogical removal of 30 per cent of the valuable convecting and radiating front panel area by introducing, in Exp. 39, an aperture 5in high and 11in long through which air tumbled in and out of the container in an uncontrolled manner. Cold air drawn in through the inlet vent collided with that sucked in at the base of the aperture, creating the turmoil illustrated by the writhing 30° isotherm below the heat source, the heat from which was transferred to the frustrating upper surface of the dust cover, with some of the hot air turning left to form the anti-clockwise internal convection pattern while the remainder turned right to escape through the upper vent, with the margins of the stream being jostled out through the top of the aperture. Ill-controlled though it was, the larger volume of cold air admitted removed a greater quantity of heat, as is shown by the increased area of the pattern outside the container and the substantial temperature reductions within. Moreover, the heat-laden air was thrown clear of a possible unit above to avoid 'chimney effect', thus meeting one of the most important requirements of rack-mounting units. Unfortunately, ventilation alone could do little towards reducing the temperature differential between upper and lower dust cover surfaces responsible for 'heat exchanger' action. This required the mobilization of additional means of heat dissipation, which were applied in a further series of experiments.

(To be continued)

A NEW TELERECORDING EQUIPMENT

Telerecording forms an important part of modern television programme techniques, for, by this means, a programme can be transferred to film for use on a subsequent occasion. Basically, this is achieved by feeding the video signals and synchronizing pulses to a cathode-ray tube capable of producing an intensely bright, short-persistence picture. This picture is optically focused on to a 'frame' of negative film stock and thereby photographed, the film then being moved on one 'frame' for the taking of the next picture, and so on. The reel of the film is then developed and processed in the normal way for future use.

Although theoretically straightforward, the practical realization of such an equipment poses many problems. For instance, in order to record full picture information, the film must be stationary in the gate for the period of the two scanning fields which make up one complete television frame or picture; this means that the only time left in which it can be moved on for the next exposure is in the brief blanking period between television frames.

As this blanking period varies between 1·4 and 1·8msec, two opposing requirements have somehow to be reconciled. Firstly, the film must (ideally) be pulled down into position in this exceedingly fast time, as any overlap into the period when the television picture is being built up by the normal scanning process will mean that some of the scanning lines (and therefore picture-content) are not registered on the film. In practice, the loss of a few lines is not of great importance; nevertheless, the nearer the pull-down time approaches these limits the better will be the quality of the filmed picture.

Secondly, it is of prime importance that the film itself must not suffer physical injury in the process of being pulled down.

Pull-down times in use on conventional interrupted cameras have been so long as to make the arrangement quite unworkable, and hitherto the usual practice has been to record only alternate television frames, the others being blanked out and the intervals used as 'breathing spaces' in which the film is pulled down.

The Marconi telerecording equipment type BD.679 incorporates an entirely new and exclusive fast pull-down mechanism which operates in the normal time of 2msec, but which can be adjusted to give a pull-down actually within the blanking period. A very light gate-pressure is employed, while positive registration of the film is achieved against a sapphire-tipped claw. The fast pull-down mechanism itself is totally enclosed in an oil-bath and can be quickly removed as a sub-unit. A very simple optical system is incorporated which ensures that there is negligible loss of contrast in the recording due to scatter.

The camera itself is supplied with magazines of 2 250ft capacity (or 2 000ft magnetic stripe), identical for feed and take-up. 'Minutes left' indicators are provided, together with 'lost loop' and 'take-up-failure' cut-outs. The gate is designed to deal with long recording sessions without attention, and splices can be passed without difficulty. A footage indicator is incorporated in the camera.

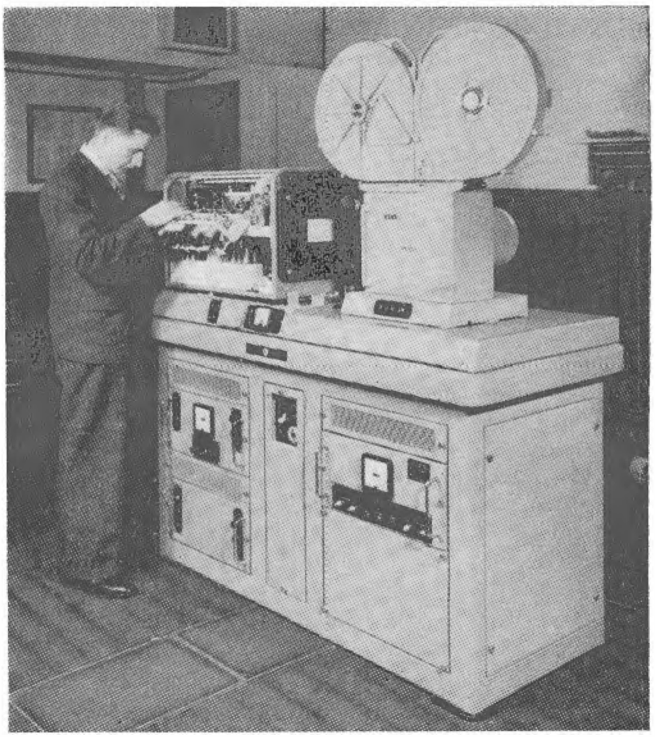
The recording monitor, the function of which is to generate (from the externally supplied video signals) an intense high-quality picture for focusing on to the film, is mounted on a swivel head, and may be swung through 90° to facilitate removal of the recorder cathode-ray tube. Due to the flexible design of this unit, any magnetically-focused tube of a diameter between 5 and 10in can be used.

A flywheel sync. panel can be supplied as an optional extra for the regeneration of line and field pulses when the recording is made from a remote location. Other optional extras are a grey scale generator, a waveform monitor and a 14in picture monitor.

One of the first of these has been purchased by a German television authority, Bayerischer Rundfunk, who sent engineers to this country to test the apparatus thoroughly before a decision was made. Installation in Germany is now completed.

The BBC has a similar equipment in use, while another is on order for an Australian television station.

A general view of the equipment with an engineer examining one of the sub-chassis from which the covers have been removed



The Characteristics of Parallel-T RC Networks

By D. H. Smith*, M.Sc.

The symmetrical parallel-T network has been studied. Typical curves of magnitude and phase of the transmission ratio are given for both balanced and unbalanced networks. An analytical treatment of the balanced network leads to an understanding of the factors affecting the selectivity. It is shown that a single stage amplifier with a balanced network in the feedback path will not in general tune to the null-point frequency.

Unbalanced networks have no null-points. For one class of unbalanced network the phase-angle is 180° at a certain frequency. With such a network connected between its anode and grid, a single stage amplifier will tune to this anti-phase frequency, and if the gain is high enough it will oscillate.

THE use of a parallel-T network as a selective circuit was first suggested by Scott[1] and his work has been followed by a number of papers on the use of such a network in rejection filters, selective amplifiers, and low frequency oscillators. Most of them, for example[2,3,4,5], deal only with networks which are capable of giving zero transmissions at a particular frequency; these are often referred to as 'balanced' filters. The use of 'unbalanced' networks has been discussed by several authors[6,7,8,9]. In this article a more general account of the characteristics of parallel-T networks is given; the way in which load and source resistances affect the selectivity is investigated, and the use of these networks in feedback circuits is discussed.

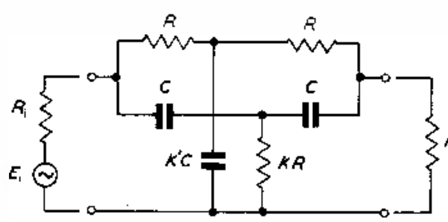


Fig. 1. Symmetrical parallel-T network

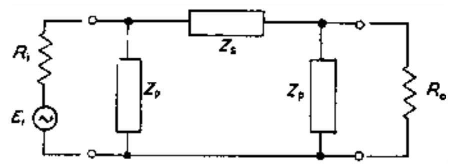


Fig. 2. Equivalent π-circuit

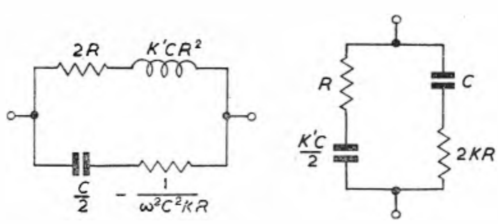


Fig. 3(a). (left) series arm Z_s, (b) (right) shunt arm Z_p

Transmission Characteristics

The most general form of the symmetrical parallel-T network is shown in Fig. 1. The unsymmetrical network, in which the series components are unequal has similar properties, and is less convenient for most applications. As Stanton[3] and Bolle[8] have shown, it is capable of producing greater selectivity.

The transmission ratio β of the network is defined by the relation

$$\beta = (E_o/E_i) \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

β is in general a complex quantity with a magnitude between 0 and 1. To evaluate β the circuit in Fig. 1 can be transformed into the equivalent π-network shown in Fig. 2. Z_s and Z_p each consist of two parallel branches as shown in Fig. 3. The total resistance in Z_s will be zero at a frequency given by the equation

$$\omega_1 = \sqrt{(1/2K)}\,.\,(1/RC) \quad \ldots\ldots\ldots\ldots \quad (2)$$

The reactance will be zero at a frequency given by

$$\omega_2 = \sqrt{(2/K')}\,.\,(1/RC) \quad \ldots\ldots\ldots\ldots \quad (3)$$

If $K' = 4K$ then $\omega_1 = \omega_2 = \omega_\infty$ say. Z_s becomes infinite and $\beta = 0$ at this frequency. In the general case $K' \neq 4K$ and at no frequency does $\beta = 0$. From Fig. 2 it can be shown that

$$\beta = \frac{1}{1+(Z_s/Z_p)[1+(R_i/R_o)]+R_i/Z_p[2+(Z_s/Z_p)]+R_i/R_o+Z_s/R_o} \quad \ldots\ldots\ldots\ldots \quad (4)$$

where $Z_p = R_p + jX_p$ and $Z_s = R_s + jX_s$.

R_p, X_p, R_s, X_s, are all functions of frequency, and of the parameters R, K, and K'. They can be evaluated from the circuits in Figs. 3(a) and (b) and shown to have the

List of Symbols

β = transmission ratio of the network = $(E_o/E_i) = |\beta| \angle\phi$
ϕ = arg β = phase-angle
E_i = source voltage
E_o = voltage appearing across load resistance
R_i = source resistance
R_o = load resistance
R = resistance of a series element of the network
C = capacitance of a series capacitor of the network
ω = angular frequency of source voltage
ω_∞ = angular frequency at which $|\beta| = 0$
ω_{min} = angular frequency at which $|\beta|$ is a minimum
ω_π, ω_0 = angular frequencies at which $\phi = \pi, 0$ respectively
$|\beta|_0, |\beta|_\infty$ = magnitudes of β at zero frequency and infinite frequency respectively
$|\beta|_\pi$ = magnitude of β at frequency for which $\phi = \pi$
K, K' are parameters defining a parallel-T network
$n = \omega RC$

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following values:

$$R_p = \frac{2KK'^2(2K+1)n^2+K'^2+8K}{K'^2(2K+1)^2n^2+(K'+2)^2} \cdot R \quad \ldots\ldots\ldots\ldots (5)$$

$$X_p = -(1/n) \cdot \frac{K'(8K^2+K')n^2+2(K'+2)}{K'^2(2K+1)^2n^2+(K'+2)^2} \cdot R \quad \ldots\ldots (6)$$

$$R_s = \frac{-KK'^2n^4+4K(2K-1)n^2+2}{K^2(2-K'n^2)^2n^2+(1-2Kn^2)^2} \cdot R \quad \ldots\ldots\ldots\ldots (7)$$

$$X_s = \frac{-2K^2K'^2n^5+4K^2(K'-2)n^3+K'n}{K^2(2-K'n^2)^2n^2+(1-2Kn^2)^2} \cdot R \quad \ldots\ldots\ldots (8)$$

where n is proportional to the frequency and equal to ωRC.

network which will be referred to later) and Bolle[8] (for the case when $R_i = 0$ and $R_o = \infty$) have derived Q-factors from analogies with series resonant circuits. It is difficult to apply these concepts to more general conditions and it is suggested that a better criterion of the selectivity is the gradient of the transmission curve at the null point. Equation (9) can then be used to find the conditions for which the gradient is a maximum.

The expression for $|\beta|$ given in equation (9) contains the parameters (R_o/R), (R_i/R) and K, but in most practical cases R_o and R_i will be fixed by conditions dependent on the external circuitry, so that in effect, R and K are the only independent parameters.

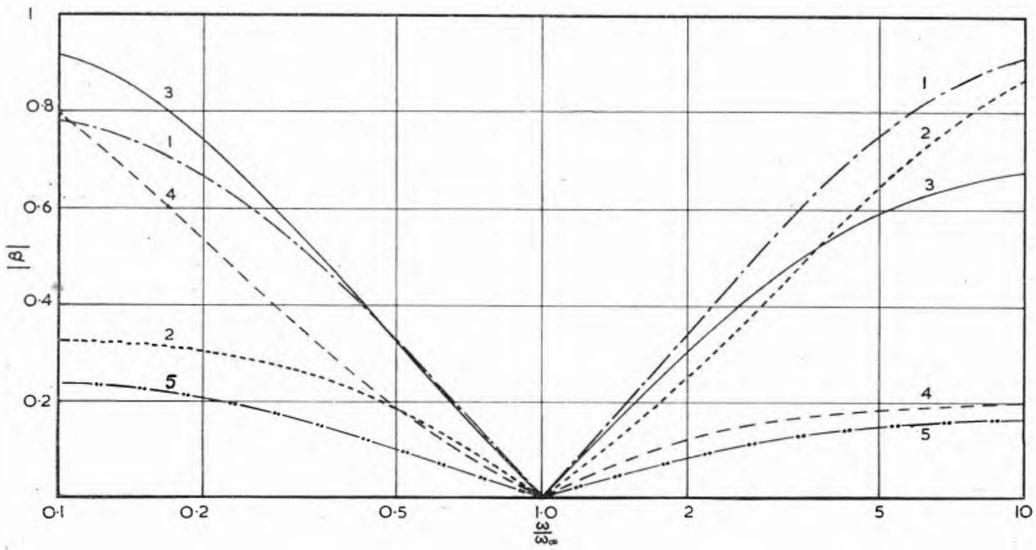


Fig. 4. Variation of transmission ratio with normalized frequency, for a balanced network, showing the effect of source and load resistances

Curves (1) $R_i = 0$, $R_o = 10R$; (2) $R_i = 0$, $R_o = R$; (3) $R_i = 0{\cdot}1R$, $R_o = \infty$; (4) $R_i = R$, $R_o = \infty$; (5) $R_i = R_o = R$

R and C are the series elements of the network, as shown in Fig. 1. It can readily be seen that equation (4) becomes unmanageable if these values are inserted as they stand, but it simplifies considerably for the special case of a balanced network and this will be dealt with first.

Transmission Ratio When $K' = 4K$

For the balanced network defined by the relationship $K' = 4K$ equation (4) reduces to

$$\beta = \frac{1-2Kn^2}{2R/R_o + [1+(R_i/R_o)]\,(1-2Kn^2) - 2n^2(2K+1)(R_i/R) + j\left\{2n(2K+1)[1+(R_i/R)+(R_i/R_o)] + 4Kn(R/R_o)\right\}} \quad \ldots\ldots (9)$$

Equation (9) can be used to draw curves of the magnitude $|\beta|$ and the phase-angle φ, of the transmission ratio, as functions of frequency, for any combination of R_o, R_i, K, R. Moreover, the equation itself yields valuable information on the effect of variation of K, R, R_i and R_o on the network characteristics, and it will be considered in some detail.

Magnitude of β

Examples of $(|\beta|, \omega)$ curves are shown in Fig. 4. The frequency has been normalized by expressing ω in terms of ω_∞, and it has been plotted on a logarithmic scale. It is obvious that the null point is unaffected by changes in R_o and R_i, and always occurs at ω_∞, i.e., when $n = [1/\sqrt{(2K)}]$. As would be expected, the steepness of the curves decreases as R_o decreases, and as R_i increases; it is noticeable, too, that the low frequencies are more affected by R_o and the high frequencies by R_i. There has been in the literature no generally accepted definition of the selectivity of parallel-T networks, though Cowles[5] (for the symmetrically-loaded

There is no simple way of finding the combination of K and R which, for any given R_o and R_i, will give the most selective filter, but differentiation with respect to K, shows that, for a given R, the optimum value of K can be obtained from the equation:

$$4K^3[R'+(R/R_o)]^2 - K[R'^2 + (2R_i^2/R^2) - 2(R_i/R_o)] - (R_i^2/R^2) = 0 \quad \ldots\ldots\ldots (10)$$

where $R' = 1 + (R_i/R) + (R_i/R_o)$.

When R, R_i, R_o, are known, it is easy to solve for K. A feature of equation (10) is that it can have only one real positive root, and this root is always between 0 and 1.

It is easy to show that if $(R_i/R) \to 0$, $K_{opt.}$ is always between 0 and 0·5, and if $(R_o/R) \to \infty$, $K_{opt.}$ is always between 0·5 and 1. The gradient at the null point is always between 0 and 0·5, the latter value being obtained only when $(R_i/R) = 0$ and $(R_o/R) = \infty$. (N.B. It can be shown from Wolf's[10] work that as the ratio of the series resistances in Fig. 1 approaches infinity, the ultimate maximum of the gradient approaches 1.)

If a convenient value of K is chosen, and equation (9) is differentiated with respect to R, then it is found that the gradient at the null point is greatest when

$$R^2 = \frac{2K+1}{2K} \cdot R_oR_i \quad \ldots\ldots\ldots\ldots (11)$$

Equation (11) suggests that R^2 should be chosen $> R_oR_i$, and if K is confined to the range 0 to 1, then R^2 must be $> 1{\cdot}5\ R_iR_o$.

The curves in Figs. 5(a) and (b) illustrate the above results. They show also that the gradient at the null point does not depend very critically on either R or K. For the

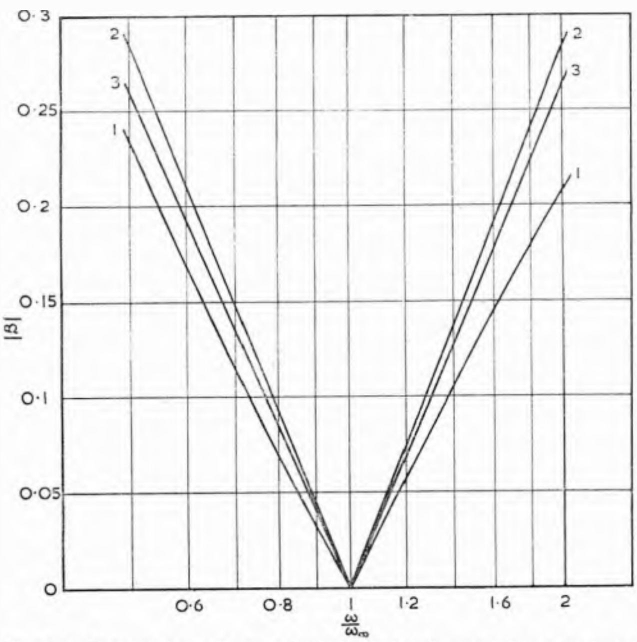


Fig. 5(a). Transmission ratio for a balanced network, showing the effect of K on the gradient at the null point

Curves (1) $K = 0{\cdot}1$; (2) $K = 0{\cdot}42$; (3) $K = 1{\cdot}0$; $R_i = 0{\cdot}1R$; $R_o = 5R$

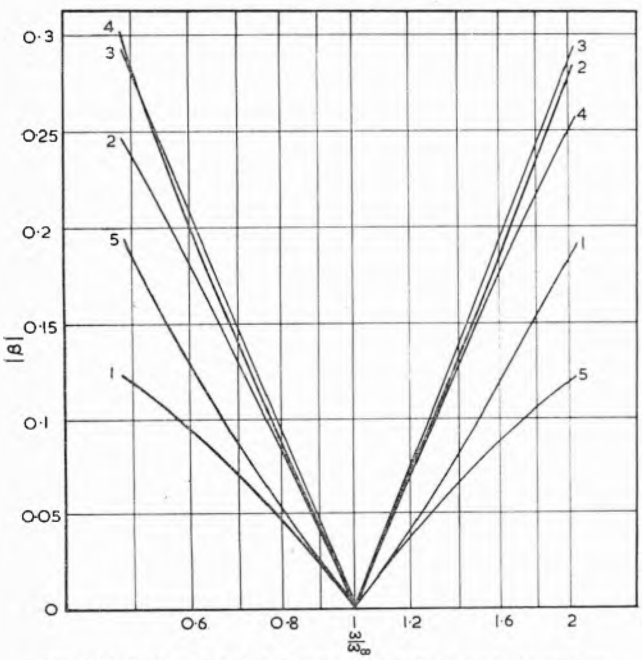


Fig. 5(b). As Fig. 5(a), but showing the effect of R on the gradient

Curves (1) $R^2 = 200R_iR_o$; (2) $R^2 = 12{\cdot}5R_iR_o$; (3) $R^2 = 2R_iR_o$; (4) $R^2 = 0{\cdot}5R_iR_o$; (5) $R^2 = 0{\cdot}02R_iR_o$; $K = 0{\cdot}5$; $R_i/R_o = 0{\cdot}02$

chosen values of R, R_i and R_o, equation (10) gives the optimum value of K as 0·42, and this value of K was used to draw curve 2 of Fig. 5(a).

It is obviously easier to choose a value of K, and use this in equation (11) to find the best value of R, than to choose R and use it in equation (10). So the best practical procedure is to choose a convenient value of K, say 0·5, and solve equation (11) to get the best value for R.

It may be noted that the gradient at the null point is limited by the ratio of R_o/R_i, e.g. when $R_o = R_i$, the highest achievable gradient is approximately 0·165 and when $R_o = 50R_i$ it is about 0·415.

The relationship given in equation (11) is particularly interesting since it corresponds to the expression derived by Cowles[5] who showed that it leads to a transmission curve which is symmetrical about ω_∞. Cowles obtained the expression by equating β_o, the transmission at zero frequency, to β_∞ the transmission as the frequency -> infinity. i.e. as can be seen from Fig. 1.

$$\beta_o = \frac{R_o}{R_o + R_i - 2R}$$

$$\beta_\infty = \frac{R_o}{R_o + R_i + (2K+1)/K \ . \ (R_iR_o/R)}$$

and if $\beta_o = \beta_\infty$, $R^2 = [(2K + 1)/2K]\ R_oR_i$, as in equation (11). When $R^2 > [(2K + 1)/2K]\ R_oR_i$, $|\beta|_o < |\beta|_\infty$ and vice versa. Curves 1 and 5 in Fig. 5(b) show the resulting contrast in asymmetry. The symbol R_{opt} will be used to denote the optimum value of R as defined by equation (11).

THE PHASE-ANGLE

The phase-angle ϕ may be of little interest when the network is being used as a rejection filter, but it is very important in a feedback circuit such as that shown in Fig. 6. If it can be assumed that there is a phase change of π radians from grid to anode of the valve, then the feedback will become regenerative for any frequencies at which ϕ is between $\pi/2$ and $3\pi/2$. If $\phi = \pi$ for any frequency ω_π, then there is a possibility of self-maintained oscillation at that frequency.

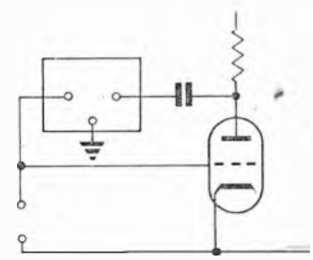


Fig. 6. Single stage feedback circuit

Equation (9) can be used to investigate the variation of ϕ with frequency, and to find the effect of the parameters K and R. Some of the results so obtained will be mentioned briefly.

$\phi \to 0$, as the frequency $\to 0$ or ∞. ϕ is negative for $\omega < \omega_\infty$ and positive for $\omega > \omega_\infty$, and there is always a phase change of 180° as ω goes through ω_∞. When $R = R_{opt}$ (or when $(R_o/R) \to \infty$ and $(R_i/R) \to 0$), $\phi = (\pm\pi/2)$ at ω_∞, and the (ϕ,ω) curve is symmetrical. In this case (curve 2 in Fig. 7) ϕ cannot exceed $\pi/2$. When $R > R_{opt}$ (curve 1 in Fig. 7) is similar but is displaced in a positive direction, so that for a limited frequency band above ω_∞, ϕ is $> (\pi/2)$. Similarly, when $R < R_{opt}$ there is a band of frequencies below ω in which $\phi < -(\pi/2)$. It is clear that when a parallel-T network is used in the feedback circuit of Fig. 6. maximum gain will not occur at ω_∞ unless $R = R_{opt}$. In the unsymmetrical case maximum gain will be obtained at the frequency at which the real part of β has the greatest negative value and, because of the regenerative effect, a higher maximum gain is possible in this case.

Transmission Ratio When $K' \neq 4K$

When $K' \neq 4K$, equation (9) does not hold and equations (4) to (8) have to be used. Fig. 8 shows the effect of the K'/K ratio on the $(|\beta|,\omega)$ curves. ω has been plotted in units of ω, since it is convenient to use the same normalizing factor for all curves having the same K, but the term ω_∞ applies strictly only to the case of $K' = 4K$.

EFFECT OF R_i AND R_o

ω_{min} is almost independent of variations in R_o and R_i, but the selectivity is affected in a manner exactly analogous

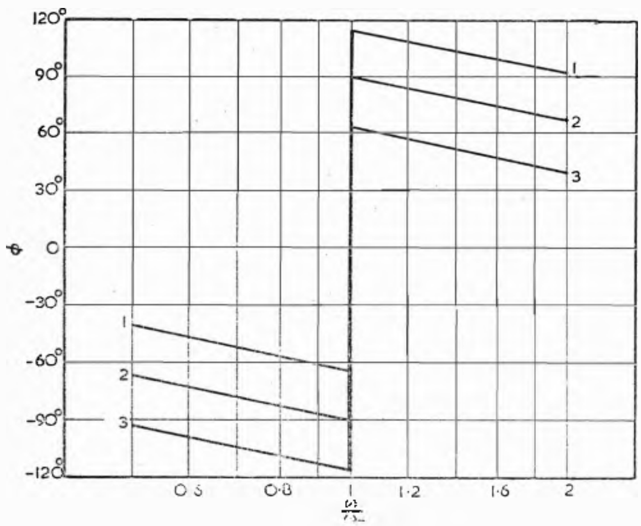


Fig. 7. Variation of phase-angle with frequency, for a balanced network, showing the effect of R

Curves (1) $R^2 = 200R_iR_o$; (2) $R^2 = 2R_iR_o$; (3) $R^2 = 0{\cdot}02R_iR_o$; $K = 0{\cdot}5$; $R_i/R_o = 0{\cdot}02$

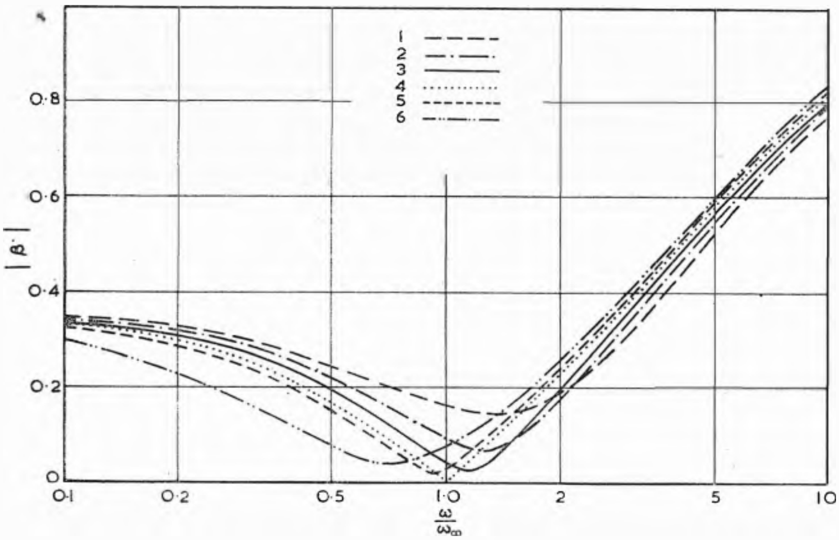
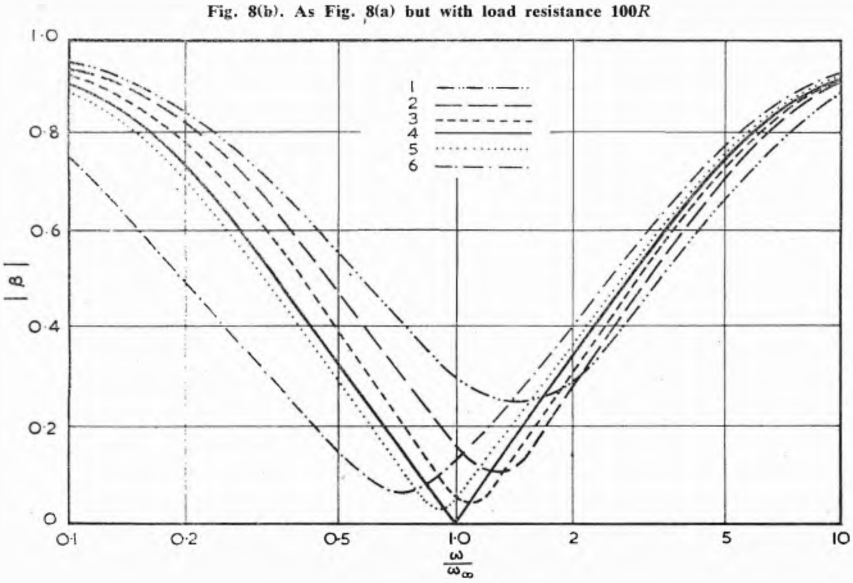


Fig. 8(a). Transmission ratio for an unbalanced network, showing the effect of the ratio K'/K
$R_i = 0$; $R_o = R$; $K = 1$; $K' = 1, 2, 3, 4, 5, 10$ *in curves* (1) *to* (6) *respectively*

Fig. 8(b). As Fig. 8(a) but with load resistance 100R



to the case when $K' = 4K$. The minima become less pronounced and $|\beta|_{min}$ decreases as the ratio R_o/R_i decreases, as a comparison of Figs. 8(a) and 8(b) shows. As one would expect, the curves are symmetrical about ω_{min} when $R = R_{opt}$.

The Phase-angle

As shown in Fig. 9 two very different types of (ϕ, ω) curve are obtained, depending on whether $K' > 4K$ or $K' < 4K$. When $K' < 4K$ the phase curve is basically of the same type as that obtained when $K' = 4K$ except that the sudden phase-change of 180° at ω_∞ no longer occurs. The magnitudes of the maxima of ϕ, and the gradient of the portion of the curve between them, both increase as K'/K increases until the limiting condition $K' = 4K$ is reached.

When $K' > 4K$, ϕ decreases continuously from 0 at very low frequencies to -2π as frequency $\rightarrow \infty$. For convenience the section of the curve between $-\pi$ and -2π is shown in the equivalent range $+\pi$ to 0, in Fig. 9. The nearer the ratio K'/K is to 4, the more closely ω_π approaches ω_∞, and the steeper is the slope of the curve at ω_π.

Also, $\omega_\pi = \omega_{min}$ when $R = R_{opt}$. For higher values of R, $\omega_\pi > \omega_{min}$ and for lower values $\omega_\pi < \omega_{min}$, but the difference is only one or two per cent even in extreme cases.

ω_π is affected by variations in K as well as by variations in K'/K. Curves of the ratio ω_π/ω_∞ as a function of K'/K for five values of K are given in Fig. 10.

In the interesting special case when $KK' = 1$, it is found that $\phi = -\pi$ (when $K' > 4K$), or $\phi = 0$ (when $K' < 4K$), always occurs when $n = 1$. Since $n = \omega RC$, this means that ω_π (or ω_o) $= 1/RC$, and so, as Bolle[8] has pointed out, if a network such as that in Fig. 1 is designed with $KK' = 1$, and the shunt elements KR and $K'C$ are varied so that their product remains equal to RC, then β is always real (i.e. $\phi = 0$ or $-\pi$) when $\omega = (1/RC)$. The magnitude of $|\beta|_\pi$ changes with variations of K, and is found to be a maximum when $K' = 4{\cdot}83$, $K \doteq 0{\cdot}207$.

It may be noted that ω_π is always $\leqslant \omega_\infty$ provided that K is regarded as fixed and K' ($> 4K$) varied. When K' is constant and K ($< (K'/4)$) is varied, ω_π is always $\geqslant \omega_\infty$.

If the network is symmetrically terminated, the (ϕ, ω) curve is symmetrical about ω_π. Unsymmetrical loading increases the slope at frequencies below and decreases the slope at frequencies above ω_π, or vice versa, according to whether R is greater or less than R_{opt}.

Input Impedance

The input impedance Z_{in} is of interest when the network is being driven by a source with an appreciable internal impedance. In general Z_{in} will be a complex function of the frequency, but as Stanton[3] has shown, when $K' = 4K$, $|Z_{in}|$ reduces to $\sqrt{[2K/(2K + 1)]}$. R at ω_∞.

When the network is connected in the feedback path of an amplifier or oscillator only a narrow band of frequencies centred on ω_π or ω_{min} is significant. In this band calculation shows that for a wide range of K, provided K' is greater than about $3K$, $|Z_s|$ is so much greater than $|Z_p|$

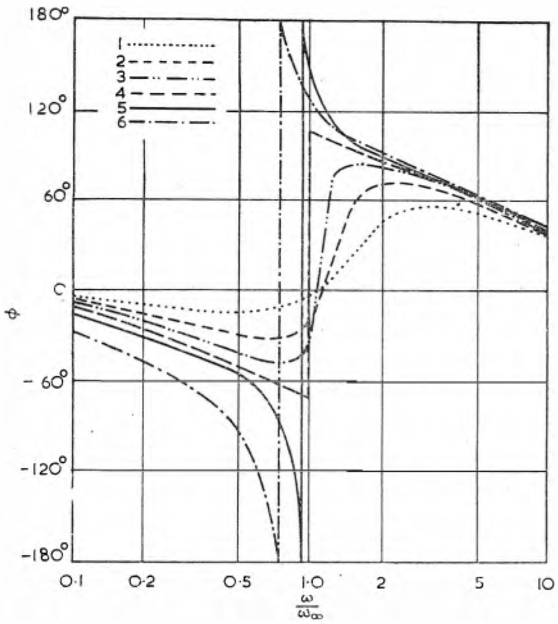
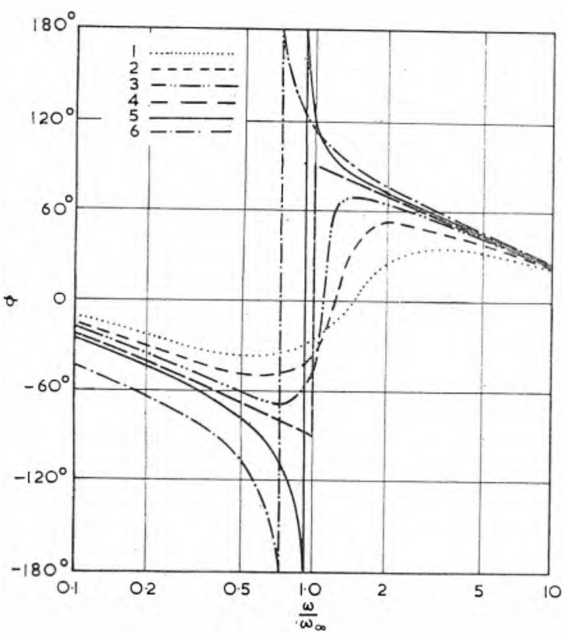


Fig. 9(a). Phase-angle for an unbalanced network, showing the effect of the ratio K'/K

$R_i = 0$; $R_o = R$; $K = 1$; $K' = 1, 2, 3, 4, 5, 10$ in curves (1) to (6) respectively

Fig. 9(b). As Fig. 9(a) but with load resistance $100R$



that $Z_{in} \simeq Z_p$. Furthermore, for frequencies within ± 20 per cent of ω_π, $|Z_p|$ does not differ by more than ± 25 per cent from its value at ω_∞. Thus for a wide range of values of K and K', the magnitude of the input impedance can be taken approximately as $\sqrt{[2K/(2K+1)]} \cdot R$.

Summary of Properties of Parallel-T Network

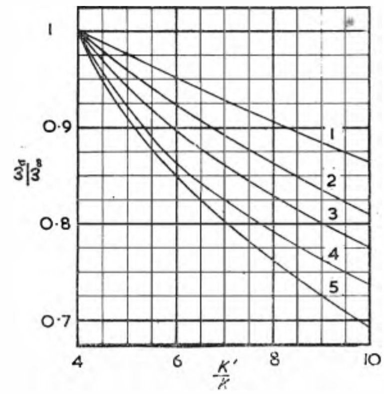
Balanced Networks

The occurrence of a null point at $\omega_\infty = \sqrt{(1/2K)} \cdot (1/RC)$ is unaffected by variations in the source and load impedances. The gradient of the amplitude-frequency curve, and hence the selectivity of the network, decreases as R_o/R_i decreases. For any fixed R_o/R_i the gradient depends on R and K, but there is no simple way of finding the best combination of these two parameters. Optimum K is always between 0 and 1 and usually close to 0·5. Optimum R is always greater than $\sqrt{(1{\cdot}5\, R_oR_i)}$ and usually less than $2\sqrt{(R_oR_i)}$. If R and K are chosen within these ranges the gradient is not critically dependent on either. When the optimum value of R, i.e. $R = R_{opt}$ is chosen, the amplitude curve is symmetrical about ω_∞.

The phase-frequency curve also is symmetrical when $R = R_{opt}$. The phase-angle then goes from zero at $\omega = 0$ to $(\pi/2)$ at ω_∞, exhibits a sudden change of π to $+(\pi/2)$ as ω goes through ω_∞ and decreases to zero at infinite frequency. For all other values of R the curve is unsymmetrical, but the phase-angle has a sudden change of π at ω_∞.

Fig. 10. Curves showing the antiphase frequency ω_π, as a function of K'/K, for several values of K

Curves (1) $K = 0{\cdot}1$; (2) $K = 0{\cdot}25$; (3) $K = 0{\cdot}5$; (4) $K = 1{\cdot}0$; (5) $K = 2{\cdot}5$



Unbalanced Networks ($K' \neq 4K$)

There is no null point in the amplitude curve, but there is always a minimum, which becomes less and less sharp as the K'/K ratio departs from 4. The selectivity is affected by R_o and R_i in the same way as in the case of the balanced filter. When $R = R_{opt}$, there is symmetry about ω_{min}.

When $K' > 4K$ the phase-angle decreases continuously from 0 at $\omega = 0$ to -2π at $\omega = \infty$. The gradient of the phase curve is greatest at ω_π, and there is symmetry about ω_π when $R = R_{opt}$; $\omega_\pi = \omega_{min}$ only for this optimum value of R. $\omega_\pi \rightarrow \omega_\infty$, and the gradient at $\omega_\pi \rightarrow \infty$, as K'/K approaches 4.

Applications of Parallel-T Networks

The parallel-T network is rarely used as a rejection filter because of its poor selectivity and its high loss at frequencies each side of the minimum. It has an important application when it is desired to attenuate very severely a single frequency (e.g. mains ripple frequency) far removed from the pass-band of a selective amplifier. However, because of the difficulty of designing LC tuned circuits at low audio and sub-audio frequencies, parallel-T networks in feedback circuits such as that in Fig. 6 are extremely useful both as oscillators and as selective amplifiers.

Parallel-T Oscillators

For the circuit in Fig. 6 to oscillate there must be a frequency at which the feedback is in phase with the grid voltage. If a phase change of π occurs in the valve, oscillation is possible only at the frequency at which $\phi = \pi$. Hence a network with $K' > 4K$ must be used, and the circuit will oscillate if the stage gain in the absence of feedback, $A > (1/|\beta|_\pi)$.

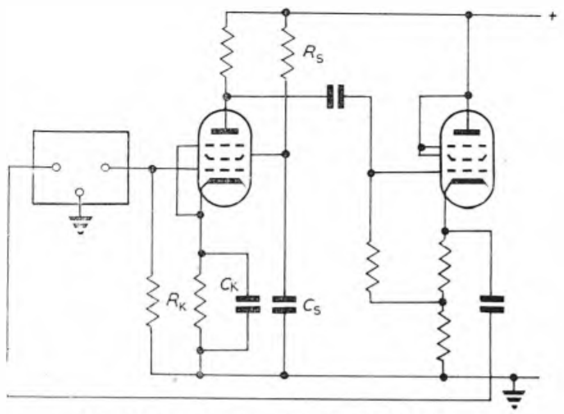


Fig. 11. Feedback circuit with cathode-follower

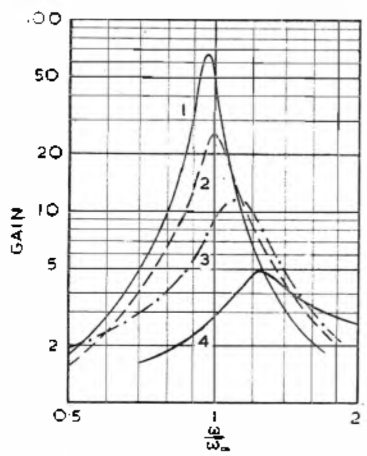


Fig. 12 Typical response curves for feedback amplifiers

Curves (1) $K' = 5K$; (2) $K' = 4K$; (3) $K' = 3K$; (4) $K' = 2K$

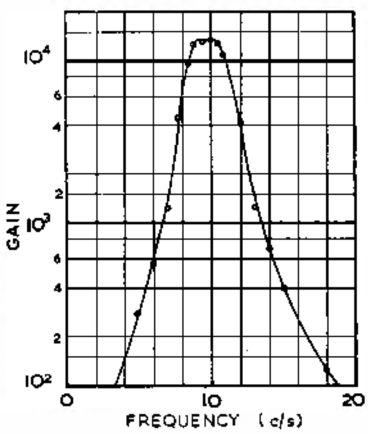


Fig. 13(a). Response curve of three-stage bandpass amplifier

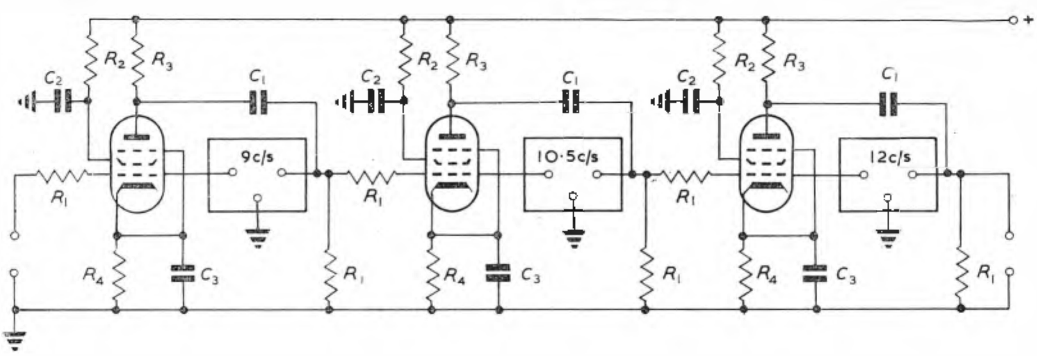


Fig. 13(b). Three-stage bandpass amplifier

Component values $R_1 = 1M$, $R_2 = 0{\cdot}5M$; $R_3 = 50K$; $R_4 = 1K$; $C_1 = 1\mu F$; $C_2 = 8\mu F$; $C_3 = 50\mu F$. Valves—3 EF37A's

For maximum frequency stability the (ϕ, ω) curve at ω_π should be as steep as possible; a given change in the phase shift in any circuit external to the parallel-T network will then cause the minimum of alteration to the frequency at which the feedback voltage is in phase with the grid voltage. It is therefore desirable to choose K'/K to be only just greater than 4. But $|\beta|_\pi$ decreases towards zero as $(K'/K) \to 4$, so that a lower limit to K'/K is set by A. It will be advantageous to choose R_o/R_i as high as possible in order to make $|\beta|_\pi$ a maximum for a given K'/K. If R is chosen so that the filter is terminated symmetrically, equal positive or negative phase changes in the external circuit will cause equal changes in the oscillation frequency.

Lynch and Robertson[7] have shown that it is possible to have the parallel-T network connected effectively between zero source resistance and infinite load resistance, by inserting a cathode-follower in the feedback path (Fig. 11). This simplifies the design calculations.

If the assumption that the anode voltage is exactly π out of phase with the grid voltage is to be justified, the screen and cathode by-pass capacitors must be large, e.g. in Fig. 11, C_s, C_K may be of such values that their reactances are equal respectively to R_s, R_K, at $0{\cdot}01\omega_\pi$. Terman[11] has shown that the phase of the output voltage may be altered by as much as 50° by incomplete by-passing.

Feedback Amplifiers

The selectivity of a feedback amplifier depends on (a) the gradients of the (ϕ, ω) and $(|\beta|, \omega)$ curves and (b) the voltage gain, A, of the amplifier without feedback. It can be seen from the graphs in Figs. 4 to 10 that a very wide range of selectivity is available. Extremely selective amplifiers can be designed in either of two ways. First, if $K' = 4K$, A may be increased almost indefinitely without instability occurring, in fact it is theoretically impossible to make the circuit oscillate when $K' = 4K$. It is possible to use three cascaded stages of amplification inside the feedback loop[1]. Such an amplifier, tuning over the range 10c/s to 100c/s by means of ganged variable resistors in the parallel-T network, has been constructed in this laboratory and found to be very stable.

The second method is to choose K'/K (> 4) so that $|\beta|_\pi$ is just less than $1/A$ when the maximum amount of regeneration, without oscillation, is obtained[3].

Some typical experimental response curves of single-stage feedback amplifiers using networks of different K'/K ratios are shown in Fig. 12.

The analysis for $K' \neq K$ shows that the feedback amplifier will 'tune' to ω_π when $K' > 4K$. When $K' = 4K$ it will tune to $\omega_\simeq$ if $R^2 = [(2K + 1)/2K] R_oR_i$ and to a lower frequency if R is less.

Bandpass Amplifier

It is not possible to design a single-stage feedback amplifier to have a steep-sided, flat-topped, bandpass characteristic.

Two or more stagger-tuned feedback amplifying stages (each using a single network) can be cascaded, and a very wide range of bandpass characteristics is available. Two three-stage amplifiers of this type have been constructed, with pass-bands of 9 to 12c/s and 60 to 80c/s respectively.

The response of the lower frequency amplifier is given in Fig. 13(a) and Fig. 13(b) shows the circuit. The usefulness of such amplifiers is limited by their inherent tendency to 'ring' when excited by transients, or when transmitting signals in the form of sharp pulses.

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Step Function Pulse Technique for Ultrasonic Measurement

Until recently ultrasonic inspection and thickness measurement techniques have been confined to sections of greater thickness than 0·25in except by the use of transverse wave methods with the accompanying disadvantages. A basically new technique has accordingly been developed by the Ultrasonoscope Co. (London) Ltd., whereby sections between 0·10in and 0·25in can be directly inspected by single probe longitudinal methods.

A major limitation of the conventional systems employing a damped train of waves is that the pulse duration is long compared with the time in which it is required to receive echoes back from close ranges. In the step function pulse technique a square wave is used, the duration of which is sufficiently long for echoes generated by the leading edge to arrive back before the trailing edge of the wave affects the receiver.

The general arrangement of the probe and specimen is shown diagrammatically in Fig. 1.

An electrical step function, Fig. 2(a), at a high repetition rate, is applied to the damping block causing a square compression wave of stress to be generated by the transmitter crystal, Fig. 2(b). This stress wave travels through the bolster, through the receiver crystal, and then into the specimen under test. The square wave will be reflected from the free surface of the specimen, or from any discontinuity within it, and will travel back through the receiver crystal.

The reflected wave, Fig. 2(c), will be reversed in phase and will also be delayed in time by the period it takes the wave to travel through the specimen. The resultant stress wave (Fig. 2(d)) affecting the receiver crystal. A simple differentiating circuit is used to give a clearer indication on a cathode-ray tube, Fig. 2(e).

The distance between the spike caused by the leading edge of the square wave and that caused by its reflection will be proportional to the thickness of the specimen in practice it has been found possible to achieve an accuracy between 1 and 2 per cent.

The technique at present covers thicknesses from 0·01in to 0·25in and has two main applications in its present stage of development, namely,

(a) Thickness measurement.

(b) Flaw detection.

Accurate measurements of thickness can be readily made from one side of the sample. These measurements are not affected by internal configurations which in many cases render the resonance methods of thickness measurement unsuitable. Typical applications are:

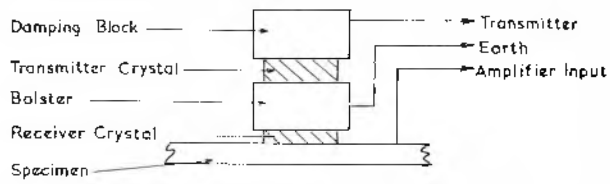
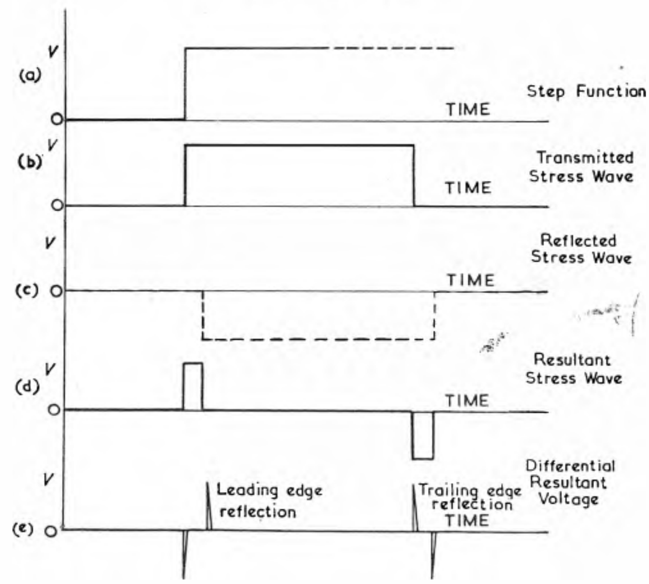


Fig. 1. The basic principle

Fig. 2. Waveforms



Measurement of distance from surface of cooling passages in turbine blades; measurements of wall thickness of jet engine combustion chambers etc. Measurement of thickness of Redux bonds, etc. Development work is under way to make it possible to utilize the technique with immersion coupling.

The method is particularly useful in delineating cracks and the laminated type of defects often found in thin sections, and is being developed for use in detecting small isolated discontinuities.

A Multiple-Channel Oscilloscope for Electrophysiology

By P. E. K. Donaldson*, M.A.

As the techniques of electrophysiology become more exact, many workers are approaching problems which can only be solved by recording simultaneously the outputs of a number of channels of information. Ideally the requisite frequency band is zero to 20kc/s (though much work can be done with a far narrower bandwidth than this) and clearly the appropriate recording apparatus is the cathode-ray oscilloscope and the photographic camera[1]. It therefore behoves the designer of electrophysiological apparatus to survey and assess the various methods of securing multiple-trace displays which are available. The writer proposes to examine the available techniques and try to evaluate them on the scores of cost per channel, complexity and performance. An instrument which embodies the principles which are felt to be the best compromise will be described.

IT is assumed that the several traces are all to appear on a single cathode-ray tube face, and for the purpose of comparison it will be assumed further: that one is aiming at an instrument with a deflexion sensitivity for each trace of the order of 50cm/V; that a time-base common to all traces will be used; that time-base generation and power supply requirements are similar for all methods and may therefore be ignored; that 'cost' is made up of two components, 'c.r.t. cost' and 'circuit cost,' and finally that the number of valves employed may be taken as a measure both of complexity and of circuit cost.

The methods of securing the multiple traces to be discussed are:

(1) Use of separate amplifying chains and electron guns for each channel.

(1a) For 2 channels only—use of separate amplifying chains for each channel, but employing only one electron gun. This is the Cossor split-beam method.

(2) The switched-input method.

(3) The oltage-coincidence method.

METHOD (1). This consists merely of repeating the apparatus for single channel recording until the required number of channels is obtained, a cathode-ray tube being used possessing a number of independent electron guns. Performance is excellent, definition and brightness maximal and possible bandwidth far in excess of that required.

Cathode-ray tubes are made in this country with one, two or four guns, and the writer understands that a three-gun tube is manufactured in France. A suitable deflexion amplifier might have three pentodes, of which two would amplify and the third supply a paraphase output for push-pull deflexion. Thus there are three valves per channel and, referring to Fig. 1, curve 1, it is seen that complexity rises linearly with the number of channels.

Unfortunately the cost does not. This is because the cathode-ray tube cost follows approximately the relation

$$\text{c.r.t. cost} \propto (\text{number of guns})^{3/2}$$

A possible cause is the greater difficulty in assembling the elaborate tubes. Another is that the proportion of rejects is liable to be greater. Whatever the reason, the cost curve rises sharply with increasing number of channels.

For the survey to be complete Method (1a) must be included here, the Cossor double beam tube which, by an ingenious arrangement, has two beams but only one gun. Where two traces only are required this is a highly economical proposition, as witness the very large number of Cossor oscilloscopes in use by physiologists both here and in the U.S.A.

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Summarizing for Method (1) then:

Performance: excellent.

Complexity: minimal and proportional to number of traces.

Cost/Trace: rises with number of traces.

Remarks: up to four traces.

METHOD (2). In this method each of a number of inputs are electronically connected in sequence to the Y deflecting

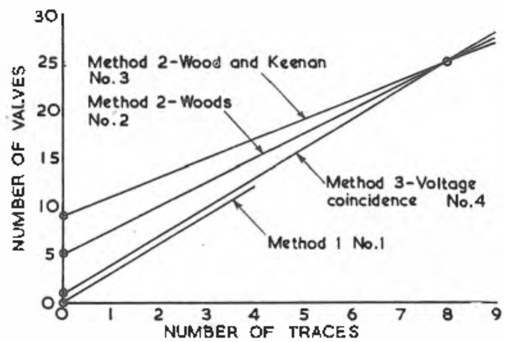

Fig. 1. Variation of circuit complexity with number of traces

plates of a cathode-ray tube having a single gun. Each input is exchanged for the next at the 'switching frequency,' so that samples of each input are obtained at a frequency:

$$\frac{\text{switching frequency}}{\text{number of channels.}}$$

Physiological electrical outputs are in general non-repetitive functions, evoked rythmically by a stimulator which triggers the time-base. It is not difficult to see that, for a function having a probability that an event will happen at any particular time, it is better to have a large number of short samples than a small number of long samples; thus one needs the highest possible sampling frequency.

The maximum usable switching frequency is usually a property of the electronic switch used and is therefore to be regarded as a constant. Thus there is a fundamental relationship for switched-input displays, that the sampling rate is inversely proportional to the number of channels, for a particular apparatus. Further, in the writer's experience, it is necessary to have at least ten samples per cycle of most waveforms if the trace is to appear satisfactory. The equation thus emerges:

$$\text{Upper limit of frequency response} \simeq \frac{\text{Maximum usable switching frequency}}{\text{Number of channels} \times 10}$$

This is the most important statement about the performance of Method (2) displays. At low switching frequencies the definition should be as good as in Method (1), and the brightness will be roughly inversely proportional to the number of traces. As the switching frequency is increased, the definition will deteriorate at each exchange of input because the beam spends an increasing proportion of its time approaching, rather than at, its correct deflexion. Definition can be restored at a further cost in brightness by gating out beam current only when the switching operation is completed.

With regard to cost and complexity, since it is possible to use a simple c.r.t., possibly a Government surplus one costing only a few shillings, we find substantially all the cost is circuit cost and therefore proportional to complexity. Taking as an example a recently-published design for a switched-input display[2], a basic complement of three valves are required (double triodes counted as one valve) and that two and a half more valves are required per channel. Since the apparatus has a gain of five, we shall need at least two more between it and the c.r.t. deflector plates, raising the basic complement to five and yielding Fig. 1, curve 2. The maximum switching frequency is 30kc/s.

In another design[3] intended for low frequencies only, nine traces are displayed. The basic complement is seven valves but only one and a half extra are needed per trace. The designers indicate that there is supposed to be an amplifier preceding each switch, and if this is taken to be half a double triode it raises the valves per trace to two. It seems probable that two more amplifying valves would be required in the common channel following the switches, raising the basic complement for the complete instrument to nine. This gives us Fig. 1, curve 3, and it is interesting to see how this design becomes economic as the proper number of channels is reached.

Summarizing for Method (2):

Performance: satisfactory, but see Fig. 2.

Complexity: more than Method (1).

Cost/Trace: falls with number of traces.

METHOD (3). This method does not seem to be widely known and, as far as the writer is aware, attention was drawn to it only very recently[1]. It is extraordinarily simple; the cathode-ray tube spot is swept vertically over the entire height of the tube face in a sinusoidal manner, so that, with a time-base also applied, a raster is formed. This sinusoidal scanning wave is also applied to a number of comparators, to each of which one input signal is fed. At the instant that the scanning wave is equal to any of the inputs, the spot is brightened for a short time. For diagrams and further explanation the reader is referred to the original article.

With regard to the performance of the method: as will be shown later, it is practicable only to use one coincidence per channel per cycle of scanning waveform. Thus, for a given scanning frequency f, the maximum frequency that can be displayed is about $f/10$, but it should be noticed that this frequency is independent of the number of traces.

It is necessary now to investigate the factors setting an upper limit to f. They are adequate brightness and adequate definition. If a 5in tube is scanned at 10kc/s, the spot moves about 10in in 100μsec, or 0·25cm/μsec. This is not a very high speed by modern standards, but is a great deal faster than would occur using the tube for physiological recording in the normal manner, and in consequence the brightness is adversely affected. A more serious difficulty lies in writing a trace which is sufficiently thin. The spot will go from top to bottom of the screen in 50μsec. A reasonable thickness for the trace is 1 per cent of the screen height, or 1/20in. The brightening pulse must then last for 0·5μsec only, and ought to have steep leading and trailing edges if the spot produced is not to be ringed by fuzz. This is about the limit of what an ordinary flip-flop will do, but nothing more elaborate can be contemplated if the instrument is to be competitive in simplicity. In the apparatus to be described, the traces are about 1/16in thick and the scanning frequency is 15kc/s. Thus the maximum frequency that can be displayed is 1·5kc/s.

Cost and complexity are again proportional because a cheap c.r.t. can be used. Referring to Fig. 1, curve 4, it will be seen that this particular voltage coincidence design has a basic complement of only one valve, and that six valves are required per pair of channels. With up to eight traces it is simpler than the examples quoted of Method (2) displays, and with above eight it is winning easily in frequency response (Fig. 2).

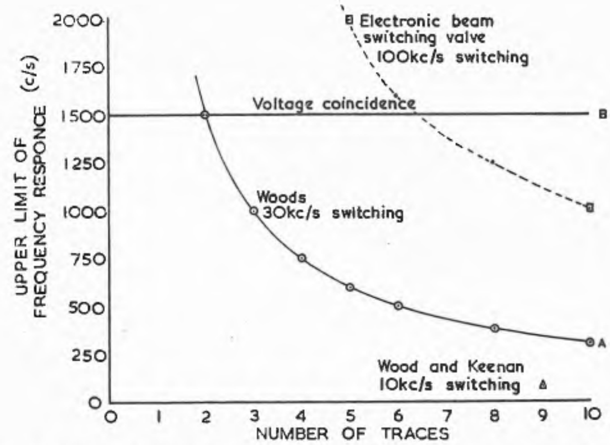

Fig. 2. Variation of upper limit of frequency response with number of traces

Summarizing for Method (3):

Performance: upper limit of frequency response of the order of 1kc/s independent of number of traces.

Complexity: compares well with Method (2).

Cost/Trace: nearly constant, lower than Method (2) below eight traces.

Appraisal

For two beams, a satisfactory solution is to do what is already so widely done, to use the Cossor split-beam tube. For four beams it is simplest and best to use a four-gun tube, but also probably most expensive. For four to six beams a switching method is indicated for a triggered time-base of less than some 10msec, in the interests of brightness. For longer and continuous running time-bases, the voltage-coincidence method is likely to be bright enough. For more than six traces the voltage-coincidence methods seems likely to hold sway unless some much faster switches appear.

As a result of the foregoing it seemed worthwhile developing a voltage-coincidence oscilloscope with initially four traces and provision for extension to eight. This was done, and it was agreeable to find that the complexity could be kept down to a satisfactory level. Some collected performance figures for each trace are:

Input resistance:	0·5MΩ
Sensitivity:	70cm/V max.
Frequency range:	zero to 1·5kc/s.
Amplitude distortion:	except at the very edge of the raster, inappreciable.

Time base:	1sec to 5msec, triggered or repetitive.
Screen diameter:	5in
Phosphor:	Green
Trace thickness:	1/16in.

It is not proposed to describe the power supplies or time-base generator, as these are quite ordinary. The tube used for this experimental model is type 5CP1, because it can be had very cheaply and has p.d.a. facilities which are useful in securing a bright trace. No work has so far been done with a tube having a proper photographic phosphor.

A description will be given of the derivation of the comparators, and the actual circuits used will then be briefly discussed.

Comparator Design

There are two functions, the fast sinusoid e_{scan} (peak-to-peak value E_{scan}) and the input signal e_{sig} (of maximum peak-to-peak value E_{sig}) and it is necessary to find means of producing a pulse of some 0·5μsec when they are equal. Let e_{sig} be applied to one end of a pair of equal resistors in series, and $-e_{scan}$ to the other; the potential of the junction, point 'O', is then zero at the required coincidence. If 'O' is connected to the grid of a valve (Fig.3(a)) where the time-constant CR is much shorter than the time of one cycle of e_{scan}, then the valve is cut off when $e_{scan} \gg e_{sig}$, is saturated when $e_{scan} \ll e_{sig}$, and produces a pip at the output of one sign or other when $e_{scan} \simeq e_{sig}$.

The usefulness of these pulses depends on the sharpness of the rising phase, and this depends on how long it takes $-e_{scan}$ to cross the grid base of the valve; the time is short if E_{scan} is large, perhaps 500V, but other considerations discussed below dictate that about a tenth of this is more appropriate. The circuit is thus quite useless as it stands; what is needed is some arrangement to take the valve rapidly from saturation to cut-off and vice-versa. Suppose positive feedback is introduced by connecting a cathode-follower from the anode to V_1 back to the cathode, we now have a Schmitt trigger (Fig. 3(b)). The circuit now snaps r rapidly between the two states and much improved pulses are obtained at the output.

By making a Schmitt trigger, backlash is introduced, which complicates matters because the 'turn-on' grid potential for V_1 is many volts greater than the 'turn off' potential. Thus when $-e_{scan}$ is positive-going ('going up'), the circuit triggers for a different value of $e_{sig} - e_{scan}$ than that at which it operates when $-e_{scan}$ is 'going down'. Setting $e_{sig} = 0$ and trying to use this arrangement to draw a flat baseline, it is found that it paints not one trace but two (Fig. 4), one above the other and separated by a distance corresponding to the backlash. One has therefore to accept a halved sampling rate of one 'coincidence' per cycle of e_{scan} only and use either 'up' coincidences or "down" coincidences. The circuit now appears as in Fig. 3(c).

It is still necessary to minimize backlash, however, in order to be able to use all the raster on the c.r.t. face. Ideally, both the range of Y-shift (obtained by adding a bias to e_{sig}) and $E_{sig(max)}$ are equal to E_{scan}. In a comparator with backlash,

$\frac{1}{2}$ shift range $= \frac{1}{2} E_{sig(max)} = \frac{1}{2} E_{scan}$ − backlash (Figs. 5 and 6).

It is undesirable to sacrifice loop gain in the trigger in an effort to reduce the backlash, as this would spoil the output pips. Fortunately one can have both fast triggering and lowered backlash by making the trigger RC coupled (Fig. 3(d)) and choosing the time-constant R_2C_2 carefully. Suppose that the common-cathode potential is zero and that V_1 cuts off at $e_{g-c} = -5$V and is saturated at $e_{g-c} = +2$V. Let $e_{sig} = 0$, so that V_1 grid potential is $-e_{scan}$. Consider the instant when $-\frac{1}{2}e_{scan}$ has just crossed, positive going, into the grid base of V_1, i.e., is -5V. V_1 anode potential begins to fall, taking V_2 grid with it. The common-cathode potential goes negative and trigger action

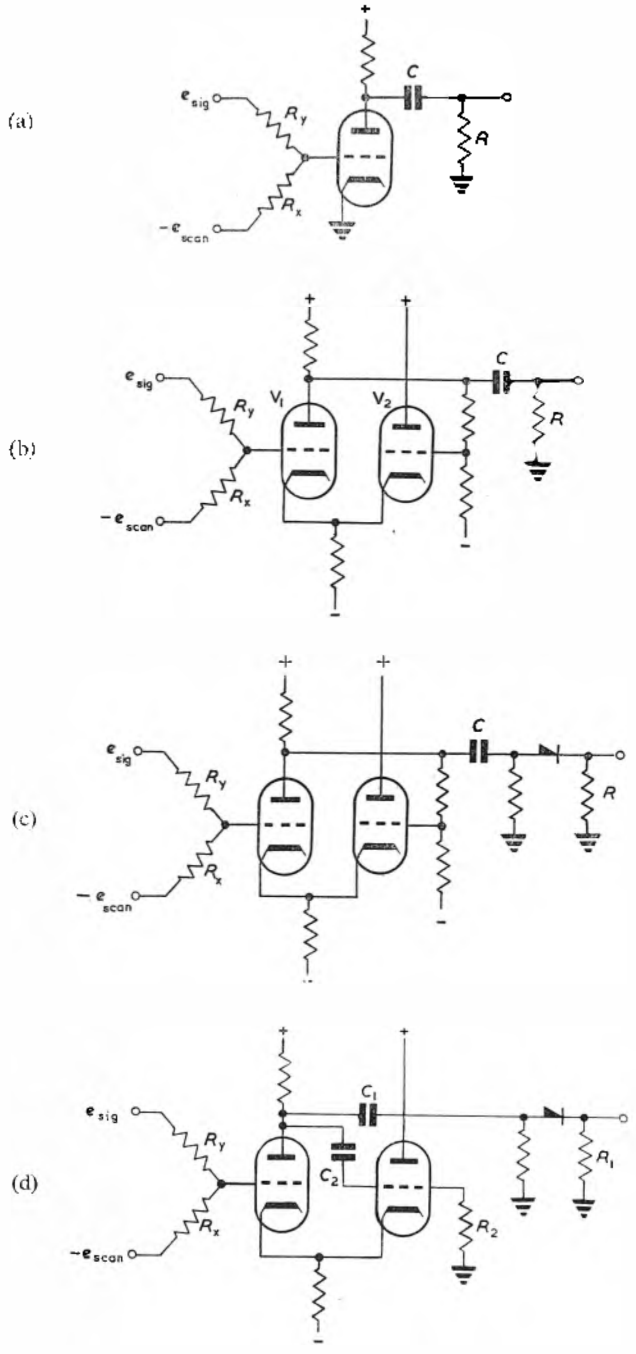

Fig. 3. Stages in the derivation of the comparator circuit

occurs. On completion of the trigger action the common cathode potential might be found to have been depressed to, perhaps, -20V, and if this were a direct-coupled trigger would remain there until V_1 grid potential fell to -18V, when reverse triggering would occur. The backlash would be $(-5) - (-18)$ or $+ 13$V.

Because of the RC coupling, however, current flows from earth to the lower plate of C_2 and V_2 grid potential rises toward earth. The common cathode potential follows it, If this time-constant C_2R_2 is so short that the common cathode potential has almost completely returned to earth before reverse triggering has to occur, then this reverse triggering hapens when V_1 grid potential is 2V. The backlash is now $(-5) - (+2)$ or -7V.

The author realizes that this is a highly artificial description, assuming as it does that nearly all the cathode current passes through V_2 (in fact, they are a long-tailed pair) and that 'cut-off' and 'saturation' are clearly-defined points. Nevertheless, he has good oscillographic evidence that the backlash does in fact reverse in sign

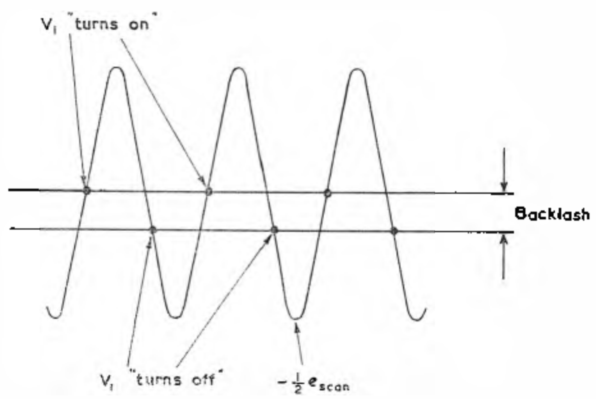

Fig. 4. Generation of double baseline when comparator has backlash, and both coincidences are used per cycle of e scan

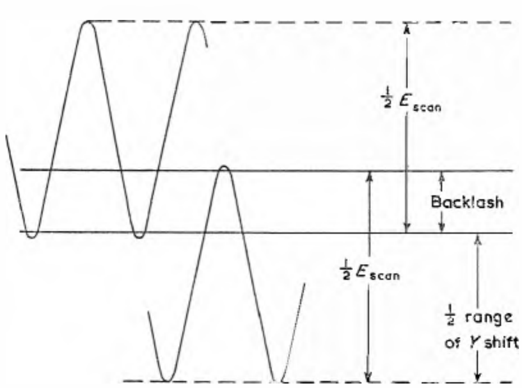

Fig. 5. Dependence of range of Y-shift on backlash

(frontlash?), and from what has gone before it is clear that this reversal is most beneficial. If the time-constant R_2C_2 is made too small, the rapidity of operation of the trigger begins to suffer; 5μsec seems about right.

With regard to the deflexion sensitivity of the system, so far it has been assumed that $R_x = R_y$, and under these conditions the sensitivity is the same as if e_{sig} were applied directly between the Y-deflector plates of the c.r.t. Making $R_x = nR_y$ (n positive, >1) the comparator triggers when $e_{scan} = ne_{sig}$; the spot is thus brightened at n times as much deflexion with the result that the apparent sensitivity of the system is increased by reducing the proportion of $-e_{scan}$ applied to V_1 grid. It does not do to try to carry the process too far, or the operation of the comparator begins to deteriorate. The writer finds $n = 10$ very satisfactory.

No combination of values of R_1 and C_1 will produce an output pulse which is large enough and crisp enough to brighten the trace directly. The pulse is therefore fed into a flip-flop which is in turn succeeded by a cathode-follower. The question then arises as to how many of these flip-flop and cathode-follower pulse-forming circuits are required. It would be very elegant if the output of all comparators could be combined in a single flip-flop, and this can be done only if the instrument is to be used to display traces which do not overlap or even approach one another very closely. The reason is that the flip-flop's 'flip' time is shorter than the pulse that initiates it. Thus the flip-flop cannot operate twice rapidly in succession because the drive pulses coalesce to produce one big one. The effect on the c.r.t. face is that, where two traces ought to cross, one disappears altogether and the other appears in the wrong place.

Nevertheless, two comparators can share one flip-flop. The method is to give V_2 in comparator number 2 an anode load, and apply to the flip-flop differentiated and rectified pulses from comparator 1, anode 1, and from comparator 2, anode 2. Comparator 1 then detects coincidences when $-e_{scan}$ is 'going up', and comparator 2 when it is 'going down'. If the resulting pair of traces are superposed, they are found to be interlaced, and no mutual interference can occur.

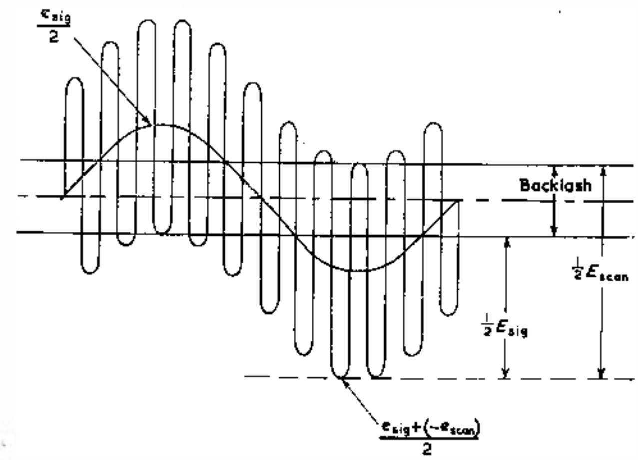

Fig. 6. Dependence of signal handling capacity on backlash

A good way to build the oscilloscope therefore is in two-channel units, each comprising two pre-amplifier valves, two comparators, one flip-flop and one cathode-follower. An indefinitely large number of these units can apparently be used together, the outputs being combined at the c.r.t. grid via crystal diodes. The circuit of such a unit is shown in Fig. 7.

The comparators are preceded by amplifying pentodes, variation of whose cathode potentials provide Y-shift. The method used is somewhat pedestrian, but has the merit that the Y-shift control has negligible effect on the gain, that maximum amplification can be had from the valve (neither of which are secured by the more usual variable cathode resistor) and that a voltmeter of very moderate 'ohms per volt' may be connected between cathode and earth to obtain calibrated Y-shift facilities if required. If a slide-back technique is used the measurement will be independent of valve gain and distortion. The input attenuator must, of course, be accurately calibrated. The +2V supply must be derived from a battery, which can conveniently be the accumulator used for the biological pre-amplifier heaters.

It will be seen that the drive from comparators to flip-flop is via potentiometers; these are necessary because the flip-flop output is appreciably dependent on the input; the potentiometers are used to equalize the flip-flop drive from the two comparators, and can be preset. The flip-flop time-constant trimmer is used to achieve a satisfactory compromise between trace thickness and brightness; the circuit has to 'flop' before it has completely finished 'flipping', hence the amplitude and duration are interrelated (Fig. 8).

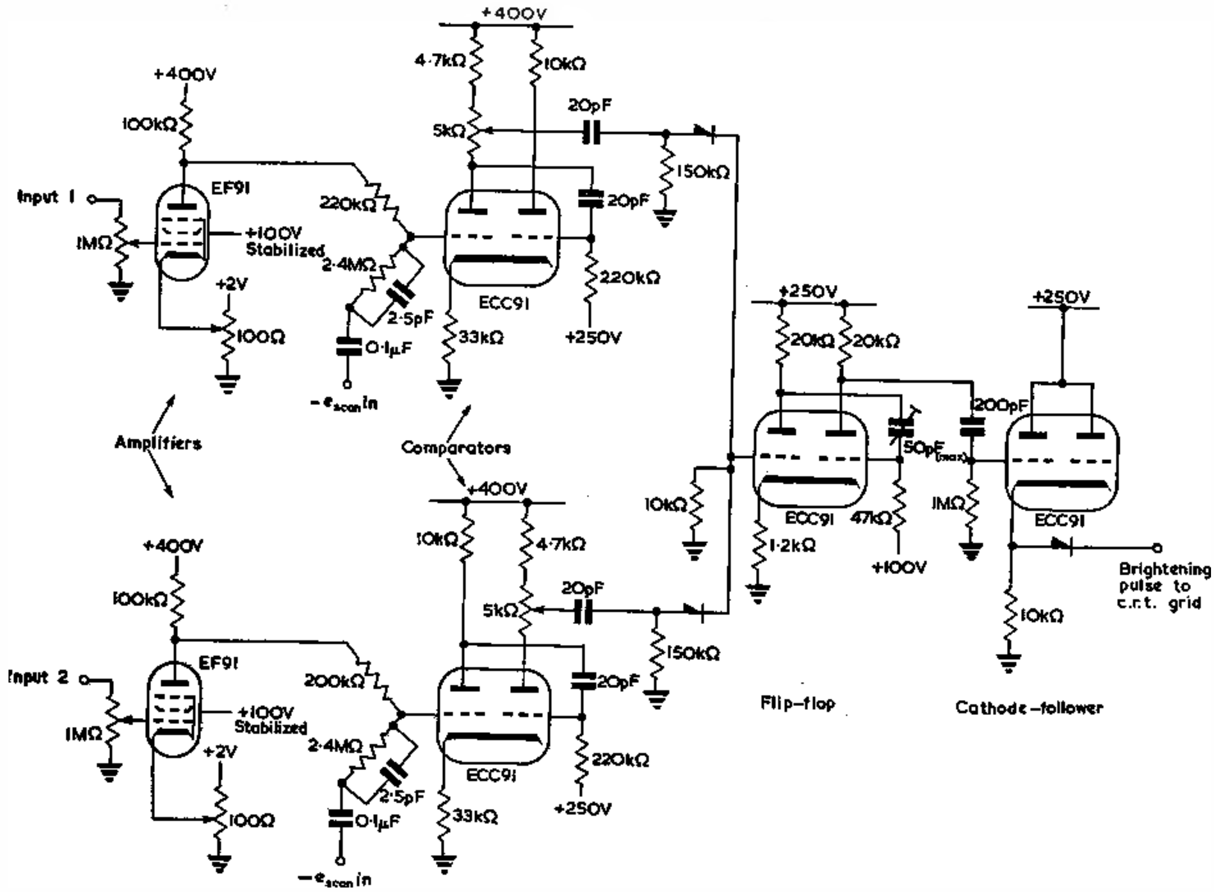

Fig. 7. Circuit of a two-channel unit

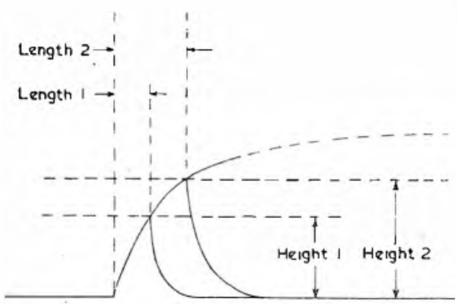

Fig. 8. Dependence of trace-brightening pulse amplitude on pulse length

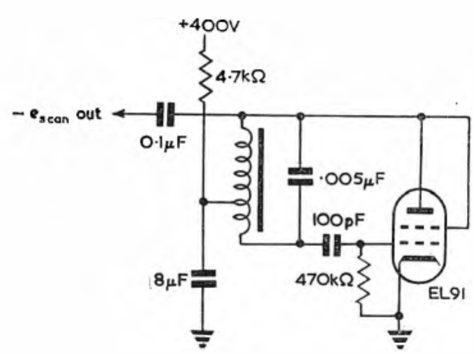

Fig. 9. The scanning oscillator

The scanning wave undergoes phase shift in passing through the 2·4MΩ resistor, unless this is shunted by a small capacitor; about 2·5pF is all that is needed. The effect of non-correction is to introduce some unwanted Y-shift. This may not matter very much, but it is well worth doing, for this reason: failure to correct means that n is frequency-conscious. Thus if e_{scan} is not a pure sine wave, but has a significant harmonic content, these harmonics will not be attenuated in the same way as the fundamental, and the fractional e_{scan} seen by the comparator has a different form from that seen by the deflector plates; the effect is to add amplitude distortion to the traces. With the correction properly carried out the oscilloscope works well with quite poor e_{scan} waveforms.

The scanning generator is shown in Fig. 9. In principle e_{scan} is applied between the deflector plates and $-e_{scan}$ to the comparators. In practice $-e_{scan}$ is applied to both and the deflector plate connexions are reversed. Only one deflector plate is fed, the other being returned to a_3. Despite this the 5CP1 shows remarkably little astigmatism or trapezium distortion. The oscillator delivers (on load to the tube and four comparators) an output of 550V peak-to-peak at 15kc/s. The waveform is surprisingly good. This output is sufficient to scan comfortably the entire face of the 5CP1 with the cathode at -2kV and the p.d.a. electrode at $+2$kV. The oscillator coil consists of 170 turns of 36 s.w.g. enamelled wire on a Ferroxcube FX.1105 core, tapped at 20 turns.

Conclusion

An attempt has been made to show that, for a multiple-trace cathode-ray tube display of four beams or more there are important economic and theoretical advantages in using the voltage-coincidence type of oscilloscope. An instrument employing this principle has been developed and described. Some idea of the performance may be had from the photographs, Fig. 10, which were taken with HPS.

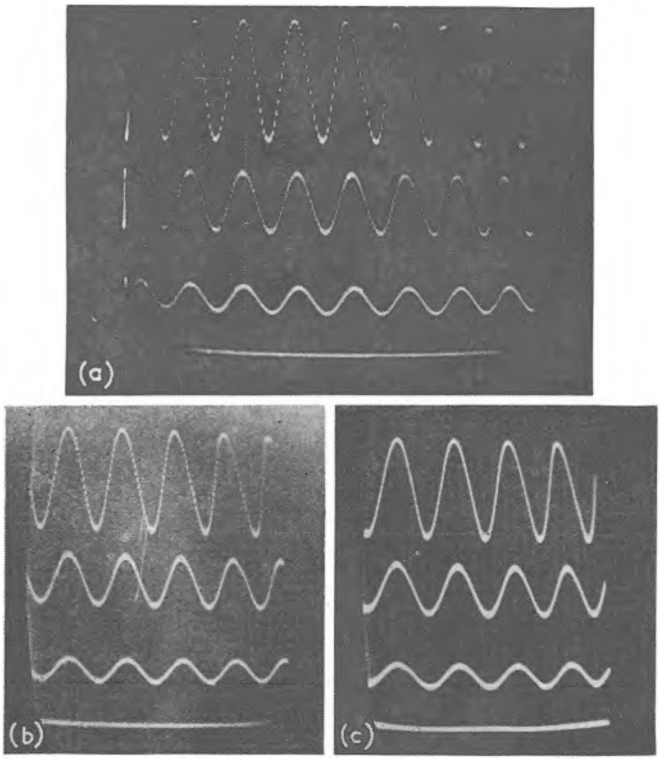

Fig. 10. Traces obtained from voltage-coincidence oscilloscope
(a) Repetitive time-base, 500c/s; (b) triggered single shot, 15msec, 300c/s; (c) triggered single shot, 50msec, 100c/s

film in a camera stopped to f/3·5. Distortion due to the curvature of the c.r.t. face is evident. In the triggered shots the time-base is short because the left-hand end was masked off to hide the resting traces. It is hoped to add beam blanking to the oscilloscope in the future.

Acknowledgment

The author thanks Mr. R. L. Gregory, of the Psychological Laboratory, Cambridge, for taking the photographs and for his valuable criticisms of the script.

ADDENDUM.

Since this article was prepared, some details have been published of a new electronic switch[5,6]. This is a special magnetron valve in which an electron beam is directed serially at one of ten anodes. A switching frequency of 100kc/s is mentioned in connexion with a commercial apparatus employing this valve. Using the criteria given for Method (2) displays, this valve gives an upper frequency response of 1kc/s for ten beams, or, by strapping the anodes in two groups of five, 2kc/s for five beams.

The author has been unable to elicit any reply from the manufacturers of this valve as to whether it is available in this country.

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The Junction Transistor as a Computing Element

(Part 2)

By E. Wolfendale*, B.Sc., A.M.I.E.E., L. P. Morgan*, B.A., and W. L. Stephenson*, B.Sc.

The Asymmetrical Bistable Circuit

The Eccles-Jordan circuit using transistors has two disadvantages. It cannot be loaded to any extent, and as the trigger points have low impedance it is difficult to trigger one circuit from another. Also the coupling capacitors introduce time-constants which restrict the maximum repetition rates.

The asymmetrical circuit of Fig. 15 was developed to overcome these restrictions. It has a simple regenerative loop, without capacitors in it, and provides a low impedance output to positive-going waveforms. The stable states are transistor 1 bottomed and transistor 1 cut-off.

The switching times of this circuit are limited by hole storage phenomena only. The simple circuit has a relatively long switch-off time due to transistor 1 being bottomed. This time can be appreciably shortened if transistor 1 is prevented from bottoming by a catching diode on the collector. Also this considerably improves the trigger sensitivity of the circuit.

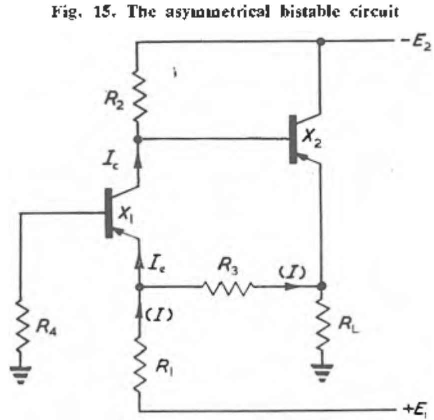
One slight disadvantage of the circuit is the need to use ± 1 per cent components to ensure that the trigger sensitivity does not vary appreciably from circuit to circuit.

* *Mullard Research Laboratories*

DESIGN OF THE BASIC CIRCUIT

Only the design of the basic circuit without catching diodes will be considered. The design of the circuit with catching diodes is complex, but an accurate design procedure has been developed. The regenerative loop will be considered first and then the necessary equations to design the stable states will be derived.

Fig. 15. The asymmetrical bistable circuit



The Regenerative Loop

An exact analysis of the loop gain over the whole frequency range involved is complex, but at low frequencies it can be shown to be approximately R_2/R_3 and if the circuit is to be bistable $R_2 > R_3$. A simple derivation of this is as follows. In the circuit of Fig. 16 the connexion between the collector of transistor 1 and the base of transistor 2 is broken and a voltage V_{in} applied between the emitter of transistor 1 and the base of transistor 2. This will cause a current of approximately V_{in}/R_3 to flow in R_3. Since transistor 1 is in the grounded base connexion (R_4 being small) its input impedance is small compared with R_1 and all the current in R_3 can be considered to flow in transistor 1. This causes a collector current of approximately V_{in}/R_3 ($\alpha \simeq 1$) and therefore an output voltage of $(V_{in}R_2)/R_3$ at the collector. The loop gain is therefore approximately R_2/R_3. At high frequencies the loop gain will fall due to the complex nature of α.

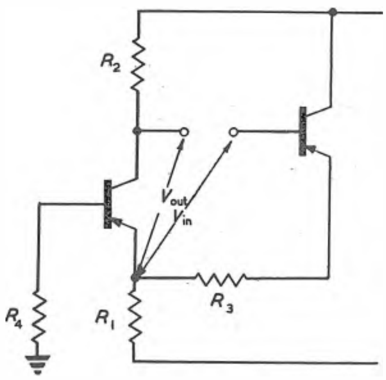


Fig. 16. The regenerative loop

The values of R_2 and R_4 must be known before the design of the circuit can be carried out and these can be fixed from considerations of pulse response and current drain of the circuit. Experiments on the circuit show that for the best pulse response R_2 and R_4 should be as small as possible, but lower limits are set by the current drain of the circuit through R_2 and the shunting of trigger pulses by R_4. Convenient values are $R_2 = 3{\cdot}3\text{k}\Omega$, $R_4 = 1\text{k}\Omega$.

Stable States

In this circuit both stable states are defined by the condition of transistor 1, it being either bottomed or cut-off. When transistor 1 is bottomed there is only a small voltage between the collector and emitter, of the order of 100mV. Therefore there is very little current through R_3 and transistor 2 and this current can be neglected in the design of the bottomed state. When transistor 1 is cut-off nearly the total supply voltage, $E_1 + E_2$, appears across $R_1 + R_3$, and R_3 is chosen to ensure that the emitter of transistor 1 is negative with respect to its base.

The bottomed state of a transistor is not well defined, as may be seen from the collector voltage/collector current characteristics of Fig. 2, in which there is no well defined knee. Therefore it is not possible to say exactly where the circuit will be stable. The bottoming action of a transistor may be considered as a process whereby the collector diode becomes biased into conduction and this leads to an alternative method of defining this stable state. The collector diode of transistor 1 is, in effect, a shunt load across R_2, so that when it conducts R_2 is shunted by a low resistance causing the loop gain to fall. The circuit will be stable if the slope resistance of the diode is less than R_3, since the loop gain must then be less than unity. The slope resistance of the diode is given by the relation $R_d = (26\text{mV}/I_d)$, so that if an approximate value of R_3 is known then a value of I_d, the forward current in the collector diode, required for stability can be calculated. In normal circuits a suitable value of R_d is $1\text{k}\Omega$, but if after designing the circuit R_3 is found to be less than $1\text{k}\Omega$ then it may be necessary to redesign the circuit using a lower value of R_d.

The design equations for the bottomed conditions are (Fig. 15):

$$I_c = \frac{E_2 + V_{cbx} + I_b R_4}{R_2} + I_d$$

where V_{cbx} is the voltage between collector and base for a collector diode forward current of I_d, this can be calculated or measured experimentally.

$$I_e = \alpha I_c = \alpha \left\{ \frac{E_2 + V_{cbx} + I_b R_4}{R_2} + I_d \right\}$$

$$V_e = V_{eb} + \left\{ (1 - \alpha) I_e + I_d \right\} R_4$$

where V_{eb} is the voltage between emitter and base for a base current of $(1 - \alpha) I_e + I_d$ and an emitter current of I_e

$$\therefore R_1 = \frac{E_2 - V_e}{I_e}$$

The bottomed state has fixed all the components except R_3 and this is used to control the cut-off state.

For transistor 2 to be cut-off, experiment shows that the emitter diode needs to be biased off by about 50mV, therefore the current in R_1 is given by

$$I = \frac{E_1 + V_{eby}}{R_1} \qquad V_{eby} = \text{cut-off bias} \simeq 50\text{mV}.$$

$$\therefore IR_3 + V_{bex2} + \left(\frac{I + I_L}{\alpha'} \right) R_2 = E_2 - V_{eby1}$$

$$R_3 = \frac{E_2 - V_{eby1} - V_{bex2} - [(I + I_L)/\alpha'] R_2}{I}$$

The equations derived here are for the limiting conditions, when the circuit will be just stable so that in practice component tolerances must be allowed for. It can be shown that for reliable operation of the circuit at high speeds it is necessary to use 1 per cent resistors and control the supplies ± 3 per cent, but in certain applications these conditions can be relaxed.

Trigger Sensitivity

The base and emitter of transistor 1 are the trigger points and as the circuit has to be triggered out of its stable state the trigger sensitivity would be expected to be very dependent of the degree of stability. This is borne out by the experimental results in Fig. 17 for the trigger sensitivity, switching transistor 1 off. The amount of trigger required increases rapidly as the base current increases, the base current being an indication of the degree of bottoming. Similar results can be obtained triggering transistor 1 on.

Typical Circuit

Fig. 18 shows a typical fully modified circuit. The choke in the collector gives an appreciable decrease in switching time. Table 3 shows the performance of the circuit for low and high frequency transistors.

Blocking Oscillators

Blocking oscillators are transformer-coupled feedback oscillators in which collector current flows for a short period, after which the transistor cuts itself off to prevent further cycles of oscillation. The pulse amplitude is limited at the collector by the transistor being bottomed. As a regenerative amplifier the blocking oscillator provides shorter rise times and pulses than with other methods of regeneration, due to the low impedance drive to the transistor.

Since the output is taken from a transformer winding, it may be taken at either polarity and at any impedance level.

TABLE 3

	L.F. (OC71) α cut-off 500kc/s	H.F. (OC45) α cut-off 5Mc/s
L	2mH	0·5mH
Minimum switching time . .	4μsec	0·5μsec
Trigger pulse . .	0.75V	0·75V

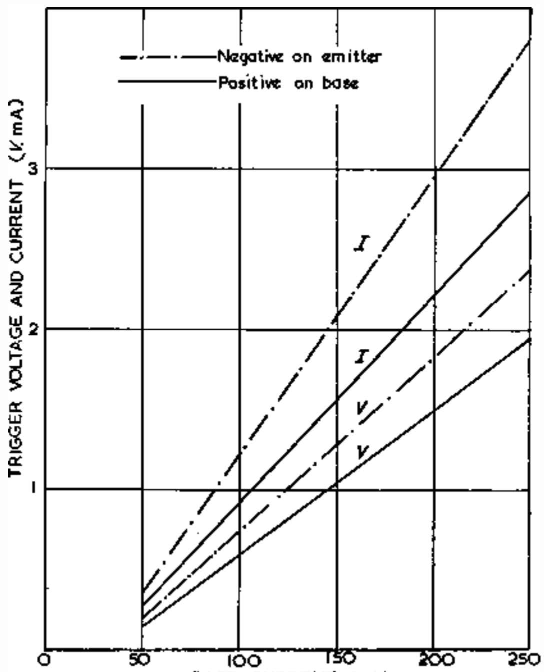


Fig. 17. Trigger sensitivity

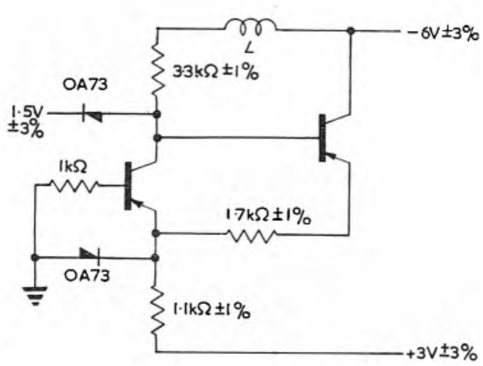


Fig. 18. Practical asymmetrical histable circuit

DESCRIPTION OF OPERATION

Fig. 19 shows the basic circuit of a blocking oscillator with feedback to the emitter, and the associated waveforms. The mode of operation is as follows:

Under no-signal conditions the transistor is almost cut-off since no bias is applied to the emitter. On the application of a trigger pulse, the transistor is brought into conduction, and regeneration causes the transistor to switch to the bottomed state. At the switching frequency, the inductive load presents a high impedance so that the collector current required to bottom the transistor is small. By means of the turns ratio of the transformer and the resistance R_1 the emitter current is arranged to be much greater than this; the excess current flows through the base.

Since the transistor is bottomed, the voltage across the primary winding of the transformer is held constant so that the collector current now starts to rise at a rate which is inversely proportional to the transformer primary inductance, the emitter current remaining constant. This rise of collector current continues until it approaches the value αI_e, whereupon the voltage across the transformer falls, and regeneration causes the transistor to be cut-off. Overshoot at the collector is damped by a diode to prevent further cycles of oscillation.

The condition for stability under no-signal conditions is that the loop gain of the circuit should be less than unity at all frequencies. This may be achieved by biasing the transistor so that the incremental emitter resistance is large enough to ensure stability. However, since the emitter resistance is very dependent on temperature, due to changes

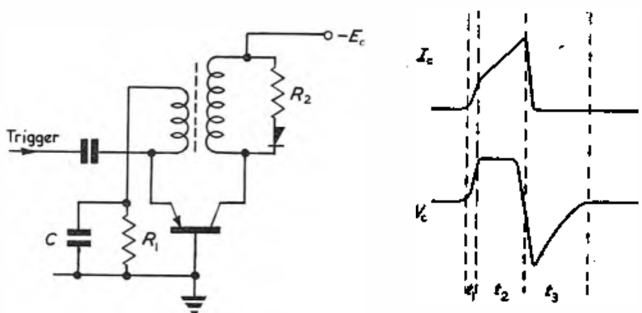


Fig. 19(a). Basic circuit of blocking oscillator
(b) Associated waveforms

in the saturation currents, it is preferable to ensure that the transistor is cut off under no-signal conditions.

Due to the presence of leakage current I_{co} and base resistance r_{bb}' stability is not necessarily assured by returning the base and emitter circuits to the same potential level. To ensure that the emitter is not biased in the forward direction, a negative bias voltage of at least $I_{co}r_{bb}'$ must be applied to the emitter. This may easily be achieved by means of a resistor between the emitter and the negative supply point, to form a potentiometer with R_1.

No attempt will be made here to give an exact analysis of the pulse shape in terms of the transistor parameters and circuit elements. However, certain relationships between pulse shape and transistor and circuit parameters have been observed, and can be used as a guide to the design of blocking oscillators for particular purposes.

Rise-Time, t_1

This is mainly dependent on the transistor parameters f_{co} and r_{bb}'. The approximate variation of rise-time with these parameters is shown in Fig. 20. The rise-time is also dependent on the transformer turns ratio since this determines the source impedance from which the transistor is driven.

Pulse length, t_2

The pulse length t_2 is given approximately by*

$$t_2 = \frac{L(\alpha n - 1)}{n^2\{R_1 + R_i\}}$$

* *The derivation of this is similar to that given by Linvill and Mattson. Proc. I.R.E. Nov. 1955 p. 1632.*

where R_i is the input resistance of the transistor, $CR_i \ll t_2$, and n is the transformer turns ratio.

Hole storage will increase this value, particularly if t_2 is small.

Reset Time

The time which must elapse before the circuit can be triggered again depends on the time-constants in the collector and the emitter circuits. These are, respectively L/R_2 and CR_1, and the longer of these two determines the reset time.

EFFECT OF LOADING

The effect of connecting a resistive load across the transformer is to reduce the current supplied to the emitter during the switching period. This increases the rise-time. The pulse width is decreased due to the greater initial step of collector current, which then rises at the same rate as before, to the same peak current.

TRIGGER SENSITIVITY

Minimum trigger requirements at the emitter are of the order of 1mA at 200mV above the latch voltage, which is the difference between the emitter bias voltage and $I_{co}r_{bb}'$. Overall trigger sensitivity is therefore dependent on temperature.

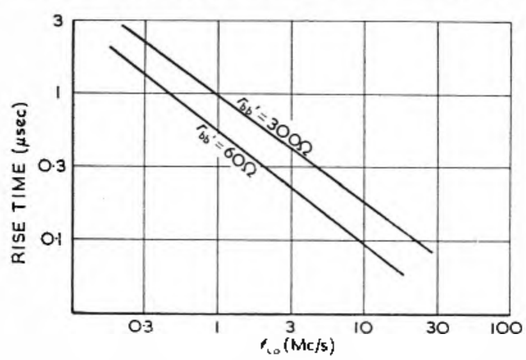


Fig. 20. Variation of rise-time with f_{co} and r_{bb}

PERFORMANCE

Typical performance figures for transistors of different types in the circuit of Fig. 19 are given below.

Circuit values: $R_1 = 330\Omega$, $R_2 = 2{\cdot}2\text{k}\Omega$, $C = 0{\cdot}01\mu\text{F}$, $E_c = 6\text{V}$.

Transformer details:

Primary: 60 turns 38 s.w.g., $L = 4{\cdot}5\text{mH}$

Secondary: 12 turns 38 s.w.g.

Core: Ferroxcube FX1011 pot core.

The emitter was biased 60mV negative by means of a 33kΩ resistor between the emitter and the negative supply.

	f_{co}(Mc/s)	$r_{bb}'(\Omega)$	$t_1(\mu\text{sec})$	$t_2(\mu\text{sec})$
OC71	0·5	300	1·8	3
OC72	0·5	60	0·8	3
OC45	5	60	0·15	2

Logical Gates

The operation of logical gates in computing systems falls roughly under two headings:

(1) Gates which pass or inhibit a pulse by means of a d.c. control potential operated from a bistable or similar circuit.

(2) Gates which show coincidence or otherwise of two pulses of similar duration, but not of controlled amplitude, which may be a.c. coupled.

In using transistors to perform these functions advantage may be taken of the amplifying and limiting properties of the transistor to provide an output which is substantially independent of the amplitude of the input signals.

D.C. CONTROLLED GATE

A circuit which will pass a pulse to one channel or another depending on the state of a bistable circuit is shown in Fig. 21. In this the control potentials at the bases of X_2 and X_3 switch in opposite phase between levels more negative than the bottoming potential of X_1 and more positive than the supply potential $-E_c$. Excursions of less than one volt are needed here in order to cut either transistor off, the collector of X_1 being held at the more negative of the base potentials. X_1 is held almost in the cut-off state by returning the base to earth, and a negative pulse is applied to its base causing it to conduct. The d.c. control levels determine whether X_2 or X_3 will conduct.

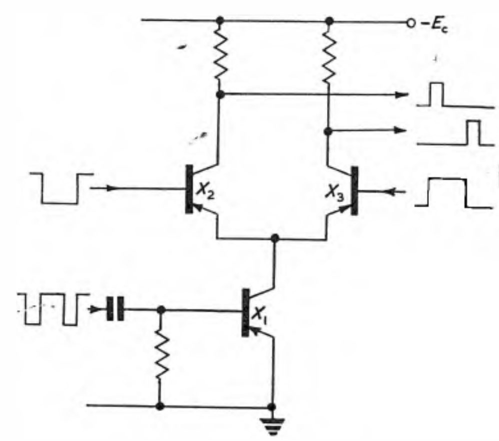


Fig. 21. D.C. controlled gate with two-channel output

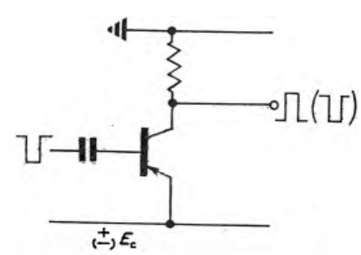


Fig. 22. Pulse inverting circuit using the symmetrical properties of the transistor

The input pulse is thus passed to one channel and not to the other. In cases where only one output is required, X_3 may be removed and a diode connected between the collector of X_1 and a potential midway between the extremes of the control voltage excursion.

Alternatively, this circuit may be extended to cover an indefinite number of output channels. A d.c. controlled circuit which uses the symmetrical properties of the transistor is shown in Fig. 22. In this the base is unbiased and the d.c. control is switched between levels which are positive and negative to earth.

When the control potential is positive, and a negative pulse is applied to the base the transistor acts as a grounded emitter amplifier, a positive output being obtained.

When the control potential is negative, the transistor acts as a grounded collector amplifier with the collector acting as emitter and the emitter as collector, so that a negative output pulse is obtained.

A.C. Coupled Gates

The two main requirements for this type of gate are to show coincidence of two pulses either by providing an output or not, as required. The circuits to perform these two functions can be similar in configuration.

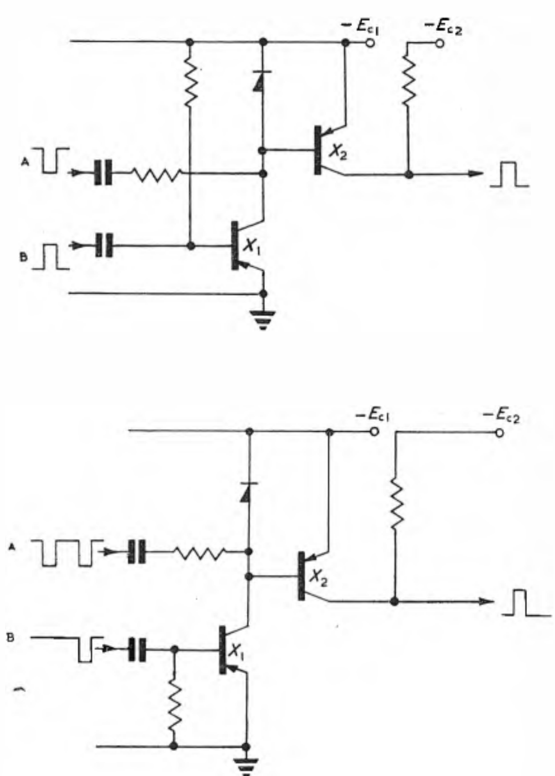


Fig. 23. Transistor biasing a diode
(a) (above) coincidence or AND gate
(b) (below) inhibit or NOT gate

Fig. 23 shows circuits which operate by biasing a diode shunt to one of the signals by means of a transistor. Fig. 23(a) shows a coincidence or AND gate in which the transistor X_1 is biased in conduction and causes the diode to form a low resistance shunt path to the applied signal A. The transistor X_2 is meanwhile cut off by the positive voltage applied to its base due to the conducting diode. On the application of a positive pulse B to the base of X_1 this transistor is cut off and the diode ceases to conduct. The applied negative pulse A is therefore applied to the base of X_2 causing this to conduct, and a positive output pulse is obtained from the collector of X_2.

If X_1 is normally cut off and a negative pulse B applied to its base (Fig. 23(b)), this circuit will act as an inhibit or NOT gate. In such a case the pulse A will pass through the circuit except when it coincides with the pulse B. The object of the transistor X_2 is to remove the pedestal voltage appearing across the diode. In some cases this may not be necessary, so that X_2 may be removed and a negative output pulse taken from the collector of X_1.

Another configuration for AND and NOT gates is shown in Fig. 24.

Fig. 24(a) shows an AND gate formed of two transistors in parallel, both biased into conduction. The load resistance and diode clamp are so arranged that the diode passes more current than either of the two transistors. Negative-going pulses are applied to the two emitters to cut off the transistors. The diode will only cease to conduct when both transistors are cut off simultaneously in which case a negative output pulse at the collectors shows coincidence of the two input signals.

Again, by arranging that one of the transistors is normally cut off instead of conducting, the circuit can be made to act as an inhibit or NOT gate, Fig. 24(b). Thus signal A will pass to the output through X_1 provided that X_2 is not conducting due to the presence of a pulse B.

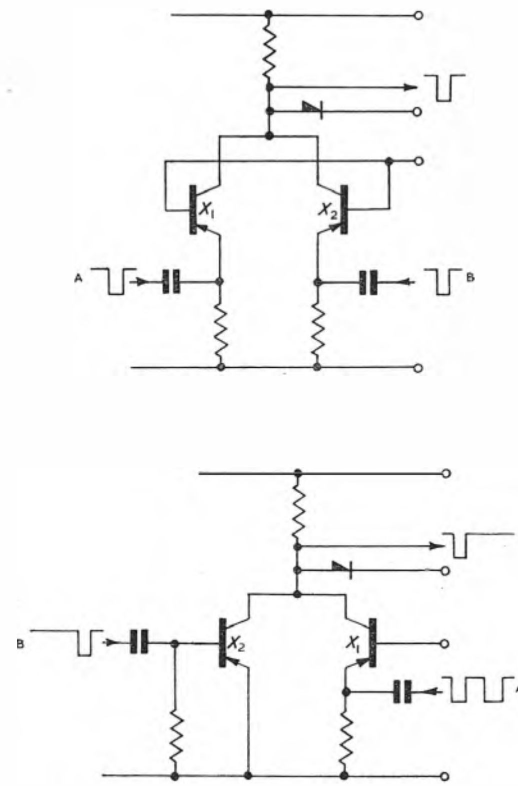


Fig. 24. Two transistor gates
(a) (above) coincidence or AND gate
(b) (below) inhibit or NOT gate

Performance

The performance of these gates depends on the hole storage characteristics of the transistors used. Referring to Fig. 7 and 8 the switching times for two types of transistor are given below.

Transistor type	OC71	OC45
f_{co}	500kc/s	5Mc/s
Turn on	6μsec	0·4μsec
Turn off: Normal conduction (Fig. 7)		
base grounded	6μsec	0·3μsec
base positive	3μsec	0·15μsec
Bottomed state (Fig. 8)		
base grounded	15μsec	0·5μsec
base positive	5μsec	0·2μsec

(To be continued)

A Stable F.M. Receiver with Preset Tuning

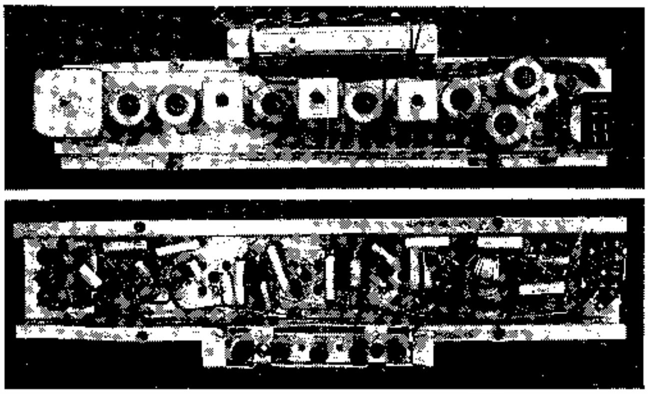
By G. S. Robinson

The physical and electrical arrangements of the receiver are described, with special reference to the use of a crystal-controlled local oscillator. A detailed description of the functions of the stages in the circuit is given, followed by a summary of their performances. Constructional details of several components are supplied.

THIS receiver was developed to obtain, as far as possible, the full benefits of an f.m. broadcast transmission, without making the usual commercial compromise between performance and economy. The requirement was for a v.h.f. adaptor with preset tuning on two or more channels in the band 85 to 95Mc/s. Because of frequency drift due to temperature effects at these frequencies, it was decided that a crystal-controlled local oscillator was desirable. The other method of ensuring stability in the absence of variable tuning, automatic frequency correction, was considered to be less economical in space, components, and complexity. A separate limiter and phase discriminator are employed, rather than a ratio detector for the reasons stated in a previous article[1], namely:

(1) More effective amplitude suppression for both $\pm\Delta f$.

(2) More effective automatic gain characteristics.



Top and underneath layout of chassis

RANGE

The performance figures show that the receiver is suitable for use in fringe areas, since the stated limit to the service area is 250μV; on the other hand, one i.f. stage could be omitted in strong signal areas.

STAGES

The receiver comprises the following seven stages, and two crystal diodes:

V_1	R.F. amplifier	EF91	or equivalents
V_2	Local oscillator and multiplier	EF91	
V_3	Mixer	EK90	
V_4	1st i.f. amplifier	EF91	
V_5	2nd i.f. amplifier	EF91	
V_6	Limiter	EF91	
V_7	Phase discriminator	EF91	
MR_1, MR_2	Discriminator diodes	GEX54	

CRYSTALS

Surplus Government 10XJ type are used. These have a tolerance of $\pm$0·01 per cent at 30°C with $\pm$·007 per cent deviation for $\pm$45°C variations, which is adequate for this purpose. In order to keep the frequency multiplication factor low, for gain purposes, a local oscillator frequency below the signal frequency was decided upon.

With an intermediate frequency of 10·7Mc/s, and a frequency multiplication factor of 10, crystals of

$$f_c = \frac{f_s - 10{\cdot}7}{10} \text{ are needed.}$$

For f_s = 89·1Mc/s, this gives f_c = 7840kc/s

f_s = 93·5Mc/s, this gives f_c = 8280kc/s.

If the desired crystals are not available, ones of the nearest obtainable value below may be ground to these frequencies.

CHASSIS

This is of standard 'strip' design, being 12in × 1in × 1½in. Screening between stages is carried out in the normal manner, by earthed sheets of metal soldered across the appropriate pins of each valve base, dividing the chassis into compartments.

WIRING

Normal v.h.f. practice is used here also, i.e. all r.f. leads as short as possible, and a common earth point is used for each stage. Heaters are earthed on one side and fed on the other, via air spaced chokes of approximately 9 turns 24 gauge p.v.c. insulated wire wound on a ¼in former.

SWITCHING

The original receiver employs a 5-section push-button unit, each section having two independent pairs of make-break contacts. The sections are used as follows:

(1) 89·1Mc/s, (2) 93·5Mc/s — see below for details

(3) Manual tuning (Switches f.m. off, a.m. on).

(4) 45Mc/s (Switches f.m. off, separate receiver on)

(5) Gram (Switches f.m. off, audio input to amplifier).

It is obvious that individual users may have other requirements. In this application, the various units are fed into a common audio amplifier, and a separate regulated power unit is used, the whole being housed in a console cabinet. (A rotary switch would simplify the switching, more contacts being available). The two pairs available per push-button are used as follows:

(1) H.T. on-off f.m.

(2) Crystal to oscillator grid.

The Circuit (Fig. 1)

R.F. AMPLIFIER

Apart from the normal advantages of improved signal-to-noise ratio and second channel rejection, V_1 eliminates the danger of radiation of local oscillator frequencies from the aerial. A low impedance input of 50Ω is obtained by means of the primary L_1 of the r.f. input transformer, which is inter-wound with L_2 on an iron dust core. The anode coil, L_3, is air spaced, and both grid and anode circuits are tuned by self and inter-electrode capacitances to midband (91·3Mc/s).

Local Oscillator and Multiplier

V_2 operates as a Pierce-Colpitts crystal oscillator with the screen grid as anode, earthed to r.f. by C_5. The appropriate crystal is switched on to the grid. The 10th harmonic of the oscillation frequency is derived by multiplication within V_2, the anode circuit being tuned to 10 times the midband crystal frequency (80·6Mc/s). A test point is provided by R_7 in series with R_6, the grid leak of the oscillator. A suitable microammeter connected across R_7 will read approximately 150μA if the oscillator is functioning correctly.

Mixer

The signal and oscillator frequencies are applied by *RC* coupling to g_3 and g_1 of V_3 respectively. The anode circuit contains the primary of T_1, which is tuned to the intermediate frequency, 10·7Mc/s.

1st I.F. Amplifier

The grid of this valve, V_4, is fed, via C_{14}, R_{14}, from the secondary of T_1. This short time-constant *CR* combination is used in the grid circuit to prevent blocking on strong signals, as no a.g.c. is employed, to avoid detuning effects and consequent phase distortion. Common anode-screen decoupling is used, to maintain good stability and high gain. The cathode resistor is not by-passed, to reduce the effects of changes in valve parameters and supply voltages.

2nd I.F. Amplifier

This stage V_5 is similar in its circuit and operation to V_4.

Limiter

The limiting action of this valve, V_6, is obtained by using low screen and anode potentials, and no standing bias. Thus the anode current swing is restricted on large signal inputs to the grid. The grid leak, R_{21}, has R_{22} in series with it, to provide a test point of limiter grid current.

Phase Discriminator (V_7)

This stage operates as an additional limiter, also having a grounded cathode, and low anode and screen potentials. The latter is held at a constant value by the potentiometer R_{23}, R_{25}. The discriminator is of the conventional Foster-Seeley type, whose method of operation is well known[2]. The discriminator diodes are of the germanium crystal type, and the audio output appears across R_{27}, R_{28}, their common load, i.e. between the junction of R_{27}, R_{29} and earth. R_{30} and C_{31} provide the required de-emphasis of the audio frequencies. C_{32} is the final audio output coupling capacitor.

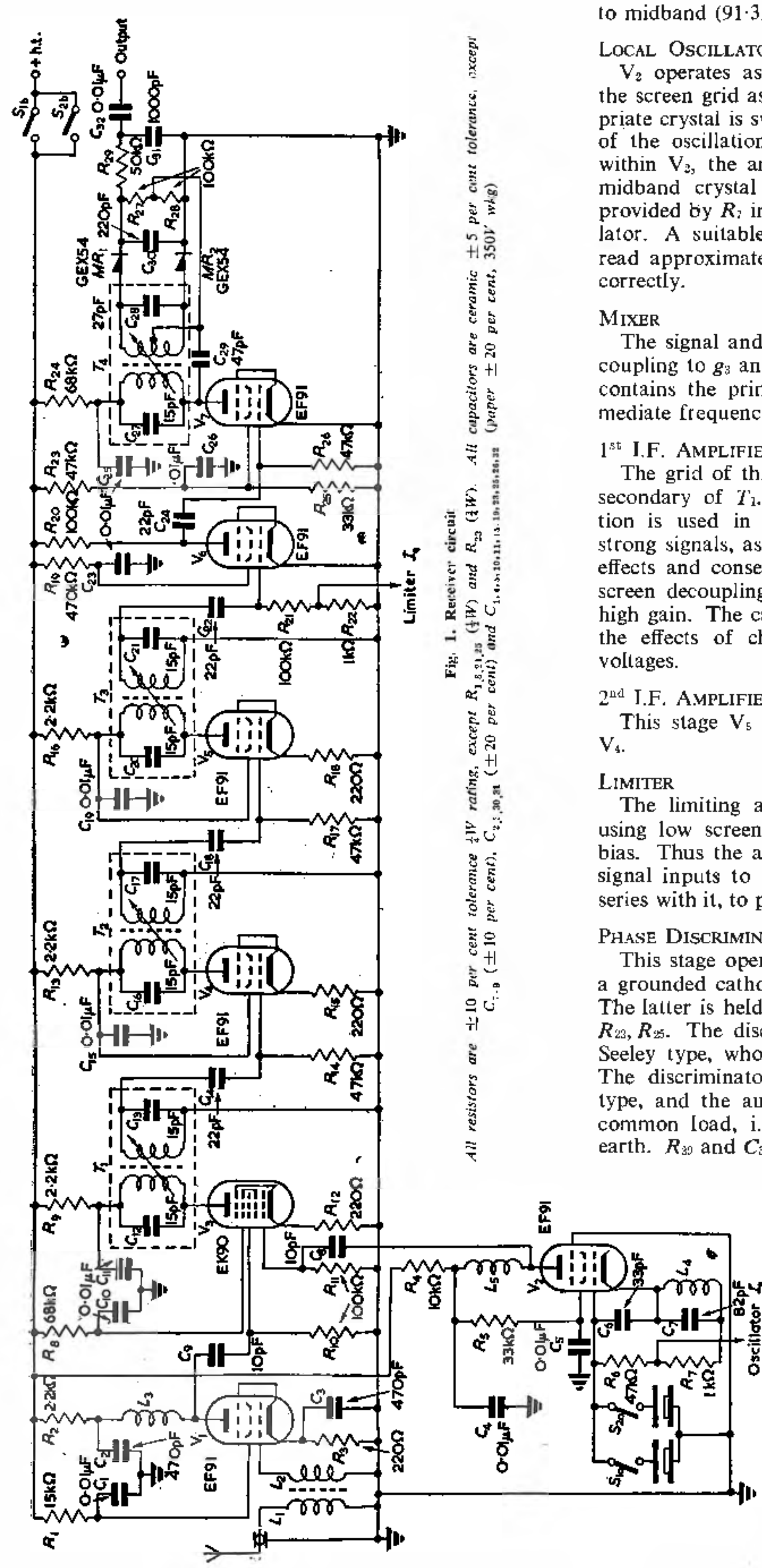


Fig. 1. Receiver circuit

All resistors are ±10 per cent tolerance ¼W rating, except $R_{1,8,21,25}$ (½W) and R_{23} (¾W). All capacitors are ceramic ±5 per cent tolerance, except C_{7-9} (±10 per cent), $C_{2,3,30,31}$ (±20 per cent) and $C_{1,4,5,10,11,15,19,23,25,26,32}$ (paper ±20 per cent, 350V wkg)

Alignment

(1) *I.F. Stages.* These are aligned in the normal way, with a microammeter connected across R_{22} to register limiter grid current. The signal generator, tuned to 10·7Mc/s (unmodulated), should be applied to the grids of V_5, V_4, V_3, in turn, reducing the input level as necessary.

(2) *Oscillator Stage.* This is tuned to midband, by opening or closing the anode coil, L_5, with non-inductive probe, to obtain maximum limiter grid current, with a midband signal (91·3Mc/s) applied to g_3 of V_3. If no

crystal is available for this frequency, adjust for equal current at both 89·1 and 93·5Mc/s input in turn.

(3) *R.F. Stage.* This is tuned to midband in the same fashion, by adjusting coils L_3 and L_2 either with an r.f. input of 91·3Mc/s at the aerial socket, for maximum, or for equal grid current at 89·1 and 93·5Mc/s in turn.

(4) *Discriminator.* This is aligned by means of a d.c. valve-voltmeter, preferably with reversible polarity, connected between the junction of R_{27}, R_{29} and chassis. (If no valve-voltmeter is available, a d.c. voltmeter with as

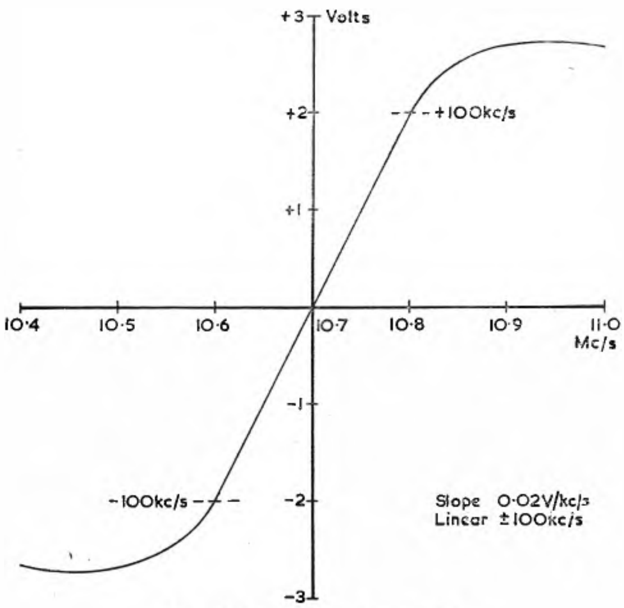


Fig. 2. Discriminator characteristic

high an input impedance as possible should be used). The procedure is as follows:

(a) With 10·7Mc/s applied to g_3 of V_3, and no crystal in circuit, adjust the secondary of T_4 for zero reading on the voltmeter.

(b) Adjust the primary of T_4 for equal peak voltages on either side of the i.f. midband. These peaks should occur at approximately 10·45 and 10·95Mc/s (see Fig. 2), and will be of opposite polarity relative to chassis.

Aerial

At a distance of approximately 50 miles from the transmitter, a standard 4-element array (dipole, reflector and 2 directors) provides approximately 150μA of limiter grid current. This is equivalent to 15V signal on the grid of V_6, ample to ensure good limiting action. The aerial is situated indoors, at about 15ft above ground level.

Performance

(1) SENSITIVITY

(a) *Overall* An audio output of 500mV r.m.s. is obtained for an r.f. input of 10μV, deviated ±75kc/s at 5kc/s.

(b) *Limiter.* The audio output attains a value of 1·5V r.m.s. when an r.f. input of 400μV deviated ±75kc/s at 5kc/s is applied. The output remains constant at this value when the r.f. input is increased beyond this level.

(2) AUDIO RESPONSE

With the de-emphasis components (R_{29}, C_{31}) out of circuit, test 1(b) produces a constant output of 1·5V r.m.s. at modulation frequencies of 400c/s, 1 kc/s, 10kc/s and 15kc/s.

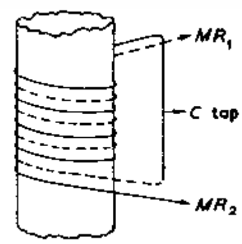


Fig. 3. Bifilar connexion of T_4

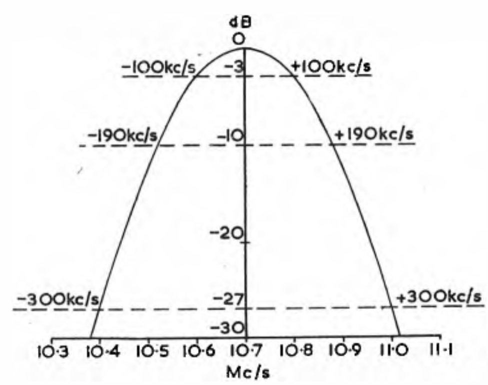


Fig. 4. I.F. response characteristic

TABLE 1
Details of Coils

DESIGNATION	WIRE	TURNS	FORMER	REMARKS
L_1	24 Gauge p.v.c.	1	¼ in dia. iron core	Inter-wound on the slug, close spaced. Self-supporting
L_2	,,	6		
L_3	,,	7	$\frac{5}{16}$in air spaced	Self-supporting
L_4	40 Gauge enamelled	200	Aladdin type PP.5948. Iron core inserted	Single pile, wave-wound. Inductance: 280μH
L_5	24 Gauge p.v.c.	8	$\frac{5}{16}$in air-spaced	Self-supporting

TABLE 2
Details of Transformers

DESIGNATION		WIRE	TURNS	FORMER	REMARKS
T_1 T_2	Primary	38 s.w.g.	38	Aladdin PP.5937	Enamelled wire
T_3	Secondary	38 s.w.g.	38	Can. No. PP.T.V.1	Spacing between windings ⅜in
T_4	Primary	34 s.w.g.	36	Aladdin PP.5937	¼in spacing between primary and secondary
	Secondary	34 s.w.g.	20 (Total)	Can. No. PP.T.V.1	Bifilar wound (close spaced) centre tapped (see Fig. 3)

(3) Second Channel Rejection

Better than 24dB at 93·5Mc/s, increasing to 28dB at 89·1Mc/s.

(4) Characteristics (Fig. 4)

(a) *Bandwidth* Within 3dB of response at 10·7Mc/s, between 10·6 and 10·8Mc/s. (Flat ±100kc/s)

(b) *Selectivity.* Response more than 26dB down, relative to 10·7Mc/s response, at ±300kc/s.

(5) Discriminator (Fig. 2)

(a) *Linearity.* Linear ±100kc/s.

(b) *Sensitivity.* 0·02V/kc/s; = 1·5V/75kc/s.

(6) Stability

By virtue of the crystal-controlled local oscillator employed, the receiver remains accurately tuned from the moment of switching on.

REFERENCES

1. Sturley, K. R. The Ratio Detector. *Wireless World 61*, 532 (1955).
2. Terman, F. E. Radio Engineering, Section 10-6 (McGraw-Hill, 1947).

Computers in Australia

SILLIAC, which began operations at Sydney University in September, 1956, is the first electronic machine of its kind to be built in Australia. UTECOM, which was set in motion one day earlier at the New South Wales University of Technology, was designed and manufactured by a British company, and assembled in Australia.

Sydney University's machine, costing £75 000 derives its name from a similar one in the United States University of Illinois.

Apart from the final assembly and wiring, carried out by University Staff, SILLIAC's construction was the work of Standard Telephones and Cables Ltd in Sydney.

UTECOM, which is slightly different in construction and has a larger memory, was produced by the English Electric Company Ltd. Installed in the University of Technology's Electrical Engineering Department, it is in great demand for theoretical work as well as commercial contracts.

It is sufficient, perhaps, to say that both are general purpose computers. However, there is some difference in their design and operating methods.

SILLIAC, which contains 2 800 valves, operates with numbers equivalent to 12 decimal digits in length. It can add or subtract groups of these numbers in 67μsec, multiply them in 667μsec and divide them in 833μsec.

Keeping up this performance as long as needed at the rate of half-a-million operations/min, it can count at the rate of a digit every 1/30μsec and add up to 1 048 576 in half a minute.

UTECOM uses numbers equivalent to nine decimal digits in length at the rate of 1 million digits/sec.

Each instruction, giving a transfer of from one to 32 numbers, takes 1/32μsec set-up time. In one hour it can solve 90 simultaneous equations with 90 unknowns. Addition requires 16μsec, while multiplication takes place at the rate of 300 operations/sec.

In a typical computation, which might run for an hour, SILLIAC can handle 30 million operations.

The data and instructions for such a problem are fed into the machine at the rate of 200 decimal digits/sec. After having performed operations on the data, the machine punches out results at 60 digits/sec or types them at 10/sec, as required.

The two universities will use the computers mainly for the many calculations associated with new developments in science and technology.

They plan also to make them available, at fees of from £50 to £80 an hour, to governmental, industrial and commercial organizations.

The Universities have now started courses of instruction in the uses of the computers for people who have problems in data processing, economics, statistics, physics, mathematics, and all fields of engineering.

An Electronically Controlled Tractor

A driverless, radio-operated tractor, the first in Australia, has been successfully demonstrated at St. Ives, near Sydney. Although only an experiment, the invention offered a glimpse into the possibilities of farming by automation.

British Farm Equipment Proprietary Limited, which represents the Ferguson Tractor Company in New South Wales and Victoria, evolved the new machine and Electric Control and Engineering Limited undertook to provide the electronic equipment for the remote tractor controls. The hydraulic equipment was supplied by C. S. Archer Pty. Ltd.

At its first public demonstration on 7 November, 1956, the tractor was controlled from a box supported by a thin spiked shaft, which was pushed into the ground like a shooting-stick.

On the control panel were eight push-buttons:

Left, Right, Slow, Fast, Up, Down, Emergency Stop, Clutch and Brake.

and from the panel, cables led to the radio transmitter and a 12V battery.

Because no provision had been made for remote starting or gear-changing, an operator started the tractor, put it into gear and walked away to the control panel.

A disk harrow was lifted clear of the ground, hydraulic pistons moved the clutch and brake simultaneously, the steering wheel swung to the right, and the tractor moved away. As it reached the edge of the demonstration plot, the harrow dropped again, biting into the soil. For two hundred yards it cut a row of straight furrows, then turned and dug another row, perfectly parallel with the first.

There was said to be no difficulty, though there would be extra expense, in fitting further equipment to start the motor and change gears. Each operation is initiated by transmission of a different radio frequency.

One possible major use for a driverless tractor is in bush-fire fighting, where such a machine could carry fire-fighting gear or a bulldozer blade right into the fire itself; and, of course, it could be used for small-scale farming in all weathers, with the operator sitting under shelter.

The inventors visualize a day when even distant operations will be remotely controlled and supervised by television.

But for the moment, they are satisfied that original research has demonstrated a simple and not unreasonably expensive robot tractor, which can be produced in quantity when the need arises.

Short News Items

Fellow Awards for 1957 announced by the Institute of Radio Engineers. Seventy-five leading radio engineers and scientists from the United States, Canada and Europe were named Fellows of the Institute of Radio Engineers by the Board of Directors at a recent meeting held in New York. The grade of Fellow is the highest membership grade offered by the IRE and is bestowed only by invitation on those who have made outstanding contributions to radio engineering or allied fields. Presentation of the awards will be made by IRE Sections all over the world wherever the recipients reside. Recognition of the awards will be made by the President of the IRE on 20 March at the Waldorf Astoria Hotel in New York during the 1957 IRE National Convention. The recipients of the award include two from England, Mr. T. H. Kinman, Manager of the Semiconductor Material Engineering Department of the British Thomson-Houston Co Ltd, and Professor F. C. Williams, Professor of Electrical Engineering, University of Manchester.

The Radio Industry Council announces that Mr. E. H. Ullrich of Standard Telephones and Cables Ltd has joined the Panel of Judges awarding the R.I.C. Premiums for Technical Writing, in place of Mr. C. E. Strong of the same firm. Mr. B. C. Brookes, lecturer on the presentation of technical information at University College, London, is also joining the Panel whose other members now are: Professor H. E. M. Barlow, Professor of Electrical Engineering, University College, London; Mr. P. D. Canning, Chairman of the Technical Co-ordinating Committee, R.I.C.; Mr. A. Clarkson, Vice-Chairman, Public Relations Committee, R.I.C.; and Vice-Admiral J. W. S. Dorling, Director, R.I.C., Chairman of the Panel.

Marconi's Wireless Telegraph Co Ltd have received a contract for the first v.h.f. multi-channel radio-telephone system in India which has been placed by the Government of India on behalf of Western Railways. The order is for multi-channel terminal units, type HM.102, with amplifying units and multi-channel terminal units type HM.104. Two radio-telephone links will be established, one between Bhavnagar and Surat, and the other between Jamnagar and Rajkot, in the Western Railways network.

The Head Office staff of the Mullard Organization have now moved to Mullard House, Torrington Place, London, W.C.1. Telephone Langham 6633.

The Electrical and Electronics Section of the Scientific Instrument Manufacturers' Association recently celebrated the tenth anniversary of the formation of the section. It was founded in December 1946 and comprised fifteen member firms; today there are 53 primary members of the section and 30 other members have a secondary interest in the Section's activities, making a total of 83 out of the whole SIMA membership of 150 instrument firms. At a recent meeting Mr. L. A. Woodhead, who is Director and General Manager of Cossor Instruments Ltd, was elected Chairman of the Electrical and Electronics Section of SIMA. Mr. A. E. Evans, Managing Director of Evans Electroselenium Ltd, was elected Vice-Chairman.

The International Union of Pure and Applied Physics is organizing in Paris, with the sponsorship of UNESCO, an International Symposium on Physical Problems of Colour Television. This will be held from 2-6 July at the Conservatoire des Arts et Metiers. Further information may be obtained from the Secretary, Colloque International sur les Problemes de la Television en Couleurs, Conservatoire National des Arts et Metiers, 292 rue Saint Martin, Paris 3.

The International British Plastics Exhibition and Convention will be held at Olympia from 10-20 July.

The London Audio Fair will be held from 12-15 April at the Waldorf Hotel, London, W.C.2. The exhibition office is at 42 Manchester Street, London, W.1.

International Aeradio Ltd have recently completed a special air traffic control installation for Vickers-Armstrongs Ltd at their airfield at Wisley, Surrey.

The South East London Technical College, Department of Electrical Engineering and Applied Physics, announces a course of eight lectures on Transistor Physics and Transistor Applications on Thursday evenings, commencing 7 February. The fee for the course is £1 and enrolment may be made by personal attendance at the College between 6.30-7 p.m. on any evening, or by post, providing the fee is enclosed and the name and address are given, together with the name and address of employer. Further information may be obtained from Mr. C. W. Robson, Head of Department of Electrical Engineering and Applied Physics, South East London Technical College, Lewisham Way, London, S.E.4. Telephone Tideway 1421/2/3.

The Technical College, Southall, Middlesex, is holding an eight-week course in Experimental Servomechanisms, commencing on 6 February, from 7-9 p.m. The course fee is £1. Further information may be obtained from the Head of the Electrical Engineering Department, Technical College, Beaconsfield Road, Southall, Middlesex.

The Aeronautical Engineering Department of the University of Southampton plans to hold another course on Acoustics from 1-6 April. It will be on similar lines to that held last year. The course is designed for people having no preliminary knowledge of acoustics. It is intended chiefly for those in the aeronautical industry, but there is adequate material to assist all those who have to deal with the increasingly important problems of noise. The fee for the course is £21, including residence in one of the Halls of the university. Further details may be obtained from D. M. A. Mercer, Physics Department, The University, Southampton. Applications should be received not later than 1 March.

Huddersfield Technical College, Department of Electrical Engineering, is holding a course of lectures on The Use of Electronics in the Textile Industry, the last three of which will be held on 6, 13 and 20 February. The subjects of these lectures will be "Irregularity Testing", "Autocounts" and "Moisture Meters, and other Applications of Electronics to Textile Machinery". Details of all courses may be obtained from The Registrar, Huddersfield Technical College, Huddersfield.

Dr. J. T. Henderson, Principal Research Officer of the National Research Council, Ottawa, has been elected President of the Institute of Radio Engineers for 1957.

Errata. "A Time Marker for Electrophysiology", by R. H. Kay, October 1956 issue, Fig. 1(a), page 453. The 300V supply to the crystal oscillator circuit should be applied to the common connexion of the 1kΩ resistor in the anode and the 47kΩ resistor in the screen circuit of the EF91 valve and not as drawn.

"Temperature Stability of Transistor Amplifiers" by G. Stuart-Monteith, December 1956 issue, page 547, the inside parenthesis of equation 52 should read $R_f + R_L$ instead of $R_L + R_L$.

In Fig. A in the letter by D. C. Pressey which appeared on page 43 of the January issue, a resistor between the anode of the 85A2 and 0V line was omitted.

Notes from ——

NORTH AMERICA

Single Element Silicon Rectifiers

Four new production types of 1 500V single-element grown junction rectifiers featuring welded case construction and the highest operating voltage commercially available are announced by Texas Instrument Incorporated. Suitable for use in series in c.r.t. power supplies and similar high voltage circuits, these miniature rectifiers operate stably to 150°C and have forward current ratings up to 125mA.

The rectifiers are available in two types of axial and stud half-wave models. The axial models, 1N588 and 1N589, allow point-to-point wiring; stud models, 1N590 and 1N591, are designed to provide maximum heat dissipation. The stud models offer a choice of anode or cathode stud, eliminating the necessity for high voltage insulation between stud and chassis.

Multiple Phototube with Ultra-violet Response

The Tube Division, Radio Corporation of America, Harrison, New Jersey, announces a 6903 head-on type of multiplier phototube intended especially for the detection and measurement of ultra-violet radiation and also useful in other applications involving low-level radiation sources.

The tube is constructed with a fused-silica faceplate which transmits radiant energy in the ultra-violet region down to and below 2 000Å. At 2 000Å, the spectral sensitivity is more than 50 per cent of the maximum response. The spectral response covers the range from about 2 000 to 6 500Å, maximum response occurring at approximately 4 400Å.

Other design features include a semi-transparent cathode having a minimum diameter of 1⅝in, ten electrostatically focused multiplying (dynode) stages, a focusing electrode with external connexion and the capability of multiplying feeble photo-electric current produced at the cathode by a median value of 4×10^5 times.

Beam Power Valve for Television Receivers

The Tube Division of the Radio Corporation of America, has introduced the 6CD6-GA high-perveance beam power valve of the glass-octal type designed especially for use as a horizontal deflexion amplifier in television receivers. Utilizing a button stem construction in a T12 envelope, the 6CD6-GA is smaller and more compact than the 6CD-G, for which it is a direct replacement, but features a modified mount design to maintain the same high perveance and to permit operation at higher ratings.

The 6CD-GA has a maximum peak positive pulse anode voltage rating of 7kV and a maximum anode dissipation of 20W. These ratings in addition to low μ-factor, high anode current rating at low anode voltage, and a high operating ratio of anode current to screen current enable this valve in suitable circuits, to deflect fully television tubes having a deflexion angle of 90°.

Like the 6CD6-G, this new valve has a structure which provides for cool operation of both control and screen grids to minimize grid emission and also provides for maximum distribution of heat to prevent hot spots on the anode.

Silicon Power Rectifiers

The General Electric Company has announced that a series of miniature 10kW silicon rectifiers is now being mass-produced. Designated type 4JA60, the rectifiers are capable of operation in an ambient temperature range from −65 to 200°C and each occupies a volume of 1in³.

It is claimed that combinations of the new rectifiers will allow power output ratings of 1MW and above.

In most circuits the rectifiers have efficiency ratings of 99 per cent and, because of the lack of ageing effects, it is unnecessary to derate them for long life.

The four rectifiers are available in peak inverse voltage ratings of 50, 100, 200 and 300V with r.m.s. voltage ratings of 35, 70, 140 and 210V.

All of the new rectifiers have a maximum reverse current rating of 50mA at maximum p.i.v. ratings at 200°C junction temperature. Maximum allowable one-cycle surge current rating is 900A and the maximum allowable continuous current is 84A/cell.

Buffer Storage Unit

Telemeter Magnetics Inc. has recently completed production of a coincident-current magnetic core storage unit. This unit, the 1092-BU-7 has properties which make it suitable for application as a temporary store, buffer or delay unit in data processing, computing and automation systems. It has a capacity of up to 1 092 characters, each of which may be up to 7 binary digits in length. The 7 bits of each character are loaded and unloaded from the memory in parallel. The characters are introduced into the store sequentially and are immediately available at the output in the same sequence as the loading sequence. Thus the store has the feature of always being ready to deliver the earliest store character regardless of whether the total number of characters in the store is 1 or 1 092.

The storage unit is completely transistorized. No vacuum tubes are employed and all components are derated to obtain the highest possible reliability.

Minimum time for loading or unloading operation is 14μsec per character with 6μsec required to switch from a loading to an unloading operation. Such switching may take place at any time in response to load and unload sync signals. There is no fixed block length for either loading or unloading. The unit emits a 'Full' signal when its capacity has been exceeded and an 'Empty' signal after the last character has been delivered.

Combined Shunt Box and Wheatstone Bridge

This instrument facilitates electrical measurements in chemistry and physics. The CM-21A 'Poly-Functionist', introduced by the Millivac Instrument Corp., Schenectady, uses a 10kΩ, 10-turn Helipot control alternatively as a calibrated shunt for the measurement and recording of d.c. currents or as the measuring arm of an a.c. or d.c. Wheatstone bridge. When used as an a.c. bridge, the Poly-Functionist delivers a polarized d.c. output signal; it is positive if the resistance being measured is lower than the balancing value and negative if it is higher. Thus, the instrument provides the possibility of making measurements in a calibrated unbalanced bridge condition with proper discrimination between above-par and below-par a.c. resistances in the same manner in which a d.c. bridge indicates a deviation of a resistance beyond and below the balance point.

Communications System for Highway

A communications system is to be installed by the General Electric Company along Massachusetts' new east-west turnpike.

The system will link v.h.f. and microwave radio equipment to provide communication for state police patrolling the road and for highway maintenance crews.

Mobile units will be equipped with selective-calling equipment to assure privacy between the departments using the facilities and selective-calling will be used also by 14 toll booths, each of which will have a v.h.f. base station.

The turnpike will have five highway maintenance areas, all of which will have microwave units for control of their vehicles and communication with police and headquarters administrative personnel

Medical Spectrometer

Model C2900 medical spectrometer by NRD Instrument Co, St. Louis, is the Francis-Bell single channel pulse height analyser designed to count pulses in the energy peak of iodine-131 and other gamma emitters. The instrument may be used to measure the spectrum of gamma radiation reaching the detector, and counts only pulses lying within an energy band whose base line and width are selectable by front panel controls.

Thus, background counting is reduced, making the C-2900 useful in the determination of iodine uptake rate in the thyroid, for localization of metastases, for accurately assaying materials for radioactive impurities, measurements of samples for tracer chemistry and for high contrast scanning of distribution of activity in tissue.

With the proper choice of detector or transducer, the analysis of varying pulse heights produced by alpha or beta radiation or by other sources can be achieved.

LETTERS TO THE EDITOR

(We do not hold ourselves responsible for the opinions of our correspondents)

A One Valve Unit For Single To Multi-Beam Oscilloscope Conversion

DEAR SIR,—This attachment enables a single beam oscilloscope to display several signals simultaneously, by sampling each signal in turn. The unit uses an Ericsson GS10D selector tube to produce a simultaneous display of ten separate outputs from an electronic analogue computer. Each signal is sampled 1 500 times per second in the circuit shown (Fig. A), so that signals of up to 100c/s can be satisfactorily viewed.

The sensitivity of the device as at present used is such that the separation between two adjacent beams is equivalent to 15V of input signal, while the amplitude of the output signal is approximately 20 per cent of the input. Thus with a 'scope sensitivity of 3V/cm the ten beams occupy 9cm of screen.

The remainder of the circuit produces the pulses to trigger the selector tube and to brighten the oscilloscope trace. A 12AT7 twin triode is used as an astable multivibrator giving an unequal mark-to-space ratio. The shorter negative pulse triggers the tube, while the positive pulse brightens the trace.

With the multivibrator *RC* values shown, in the circuit, the tube is working near its maximum frequency of 20kc/s. At frequencies above 5kc/s the cathode voltages do not rise instantly to their steady state level. It is necessary, therefore, to connect capacitors ranging between 100pF and 1 000pF across the cathode resistors of the tube to speed up the rise time of the voltage.

At frequencies below 5kc/s, the rise-time of the cathode voltage is very fast,

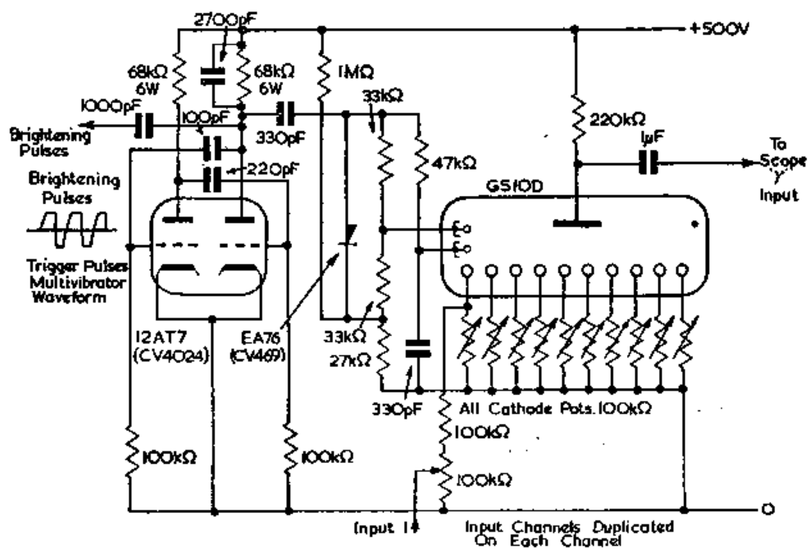


Fig. A. The multi-beam conversion circuit

The separation of the d.c. levels of the output is achieved by using different values of resistors in the ten cathodes of the GS10D selector tube. The circuit shows variable cathode resistors which enable the beams to be spaced as required for any given display. By switching out one or more of these resistors, the appropriate beams may be displaced out of sight since the d.c. level of those cathodes is then greatly different. The signal is fed to the cathode through a potentiometer and an isolating resistor. Thus, when the tube is triggered the anode samples each cathode in turn, adopting the appropriate d.c. level with the signal superimposed.

The anode of the selector tube is connected to the Y input of the oscilloscope through a 1μF capacitor, while a synchronized pulse brightens the trace during the steady state condition of the cathodes.

enabling a very clear spot to be obtained during the brightening pulse. Also, since the cathode voltage is steady when the trace is brightened, the trace may be considerably amplified without affecting the spot size to any marked extent.

There is no visible distortion of signals fed through the unit and no pulling between traces even with signal amplitudes of 100V.

It is hoped, at some future date, to use an American hard valve switching tube in place of the Ericsson tube. The American tube has a maximum frequency of 1Mc/s, thus enabling signals of up to 10kc/s to be displayed.

It is anticipated that the use of switching of information by this method will have much wider application than that mentioned in this note, particularly when several switch units synchronized by one multivibrator are employed.

The authors are grateful to Messrs. Vickers-Armstrongs (Aircraft) Limited for permission to publish this letter.

Yours faithfully,

P. W. FORTESCUE,
B. G. ATHERTON,
Weybridge,
Surrey.

A Note on Transistor Stabilizers

DEAR SIR,—At least two papers have been published recently[1,2] describing the application of transistors to stabilized supplies by virtue of low emitter to base and high collector to base impedance. When a d.c. supply is connected between collector and base and output is taken from between emitter and base then the low output impedance will confer a degree of stability to the output voltage irrespective of input or load changes. This would give a very low output voltage but this can be raised by placing a reference battery between the base and earth as shown in Fig. 1.

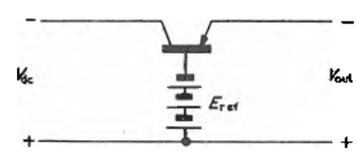


Fig. 1. Basic transistor stabilizer

The reference battery is discharged at a very low rate by the base current of the transistor. It seems to me that the logical development from here is to replace the reference battery by a stabilizer tube as shown in Fig. 2. This would have the

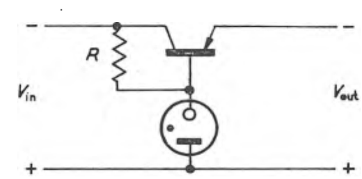


Fig. 2. Transistor stabilizer with gas filled diode stabilizer tube

following advantages over the stabilizer by itself:—

(1) This circuit will stabilize for load current changes within the current handling capacity of the transistor and irrespective of the stabilizer which has to cope only with base current and input voltage changes.

(2) An improvement results in output voltage stability against load current changes

(3) A considerable saving in input power is possible by keeping input

voltage close to the desired output voltage.

The following results given in Figs. 3 to 5 obtained with stabilizer type 85A2 and transistor type OC71 with $R = 2{\cdot}2\text{k}\Omega$ illustrate my point, showing clearly that the voltage regulation for a varying load current is better than that obtainable with an 85A2 in series with a resistor in which case a change in load current from 0 to 10mA would cause the output voltage to change by about 3V as against 0·4V obtained here (Fig. 3). Even if the supply regulation were poor (Fig. 4) the output would still vary by only 1·4V. The stability of output against input

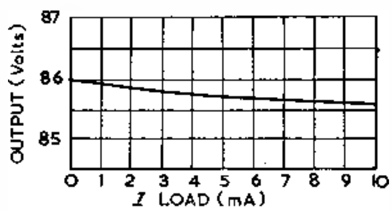


Fig. 3. Output regulation at constant input

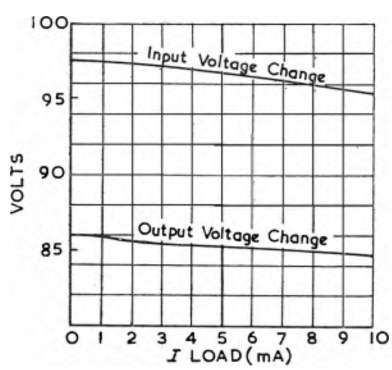


Fig. 4. Output regulation with poor input regulation

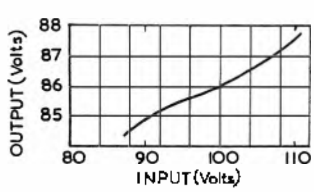


Fig. 5. Relation between input and output voltage at constant load of 5mA

voltage variation at a constant load (Fig. 5) is not as good as could be obtained in the simple stabilizer circuit with a very large value series resistor. An addition of resistance in series with the collector would improve this situation but at the cost of making the output regulation poorer. Thus the best way of utilizing this circuit might be as shown in Fig. 6. It gives voltage stability to both input and load variations. The two stabilizers must, of course, have different burning voltages.

It will be noted that the stabilizer may not strike automatically in this circuit unless R is made large, but this drawback can be overcome with a little ingenuity.

With pnp and npn transistors it would be possible to stabilize either positive or negative supply to a common earth. Manufacturers might consider the advisability of sealing in a transistor into

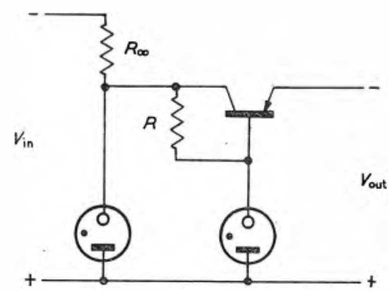


Fig. 6. Transistor stabilizer with very good stabilization characteristics

the stabilizer tube with connexions to pins.

Yours faithfully,

M. H. N. Potok,

The Royal College of Science, and Technology,
George Street,
Glasgow, C.1.

REFERENCES

1. Spencer, R. H., Gray, T. S. Transistor Voltage Regulator. *Communication and Electronics*, No 23, p. 15 (1956).
2. Keller, J. W. Regulated Transistor Power Supply Design. *Electronics 29*, 168 (Nov. 1956).

A Power Supply Stabilizer Using Transistors

Dear Sir,—Some interesting features for stabilizer circuits using transistors are illustrated by the circuit shown (Fig. 1). The first difference from vacuum-tube technique is that the series transistor X_1 is worked as a switch and is always either cut-off or bottomed. This means that the power is largely dissipated in the resistor R_1 and a small transitor can be used to handle large currents. The relatively enormous values of electrolytic capacitors available for C_1 at transistor voltages enable the inevitable ripple to be made reasonably small.

The second novel feature is that X_2 and X_3 form a d.c. super regenerative amplifier to control X_1. During 'quench' periods X_5 is cut-off and the bases of X_2 and X_3 follow the reference and output voltages respectively. When X_5 starts to conduct R_2 and R_3 provide positive feedback and the current is rapidly forced entirely into either X_2 or X_3. Thus X_1 is firmly cut-off if, and only if, the output voltage is too high. Notice that this stage contains no capacitors, since, as with ordinary super regeneratives, it is essential that there might be no memory of the previous action at the end of each 'quench' period.

The timing is controlled by X_4 and X_5 which form a free-running multivibrator of the type due to G. G. Scarott. The unequal emitter resistors R_4 and R_5 together with C_2, make a current of about 1.5mA flow through X_4 for roughly 25µsec and X_5 for 500µsec alternatively, forming the 'quench' and 'action' periods respectively.

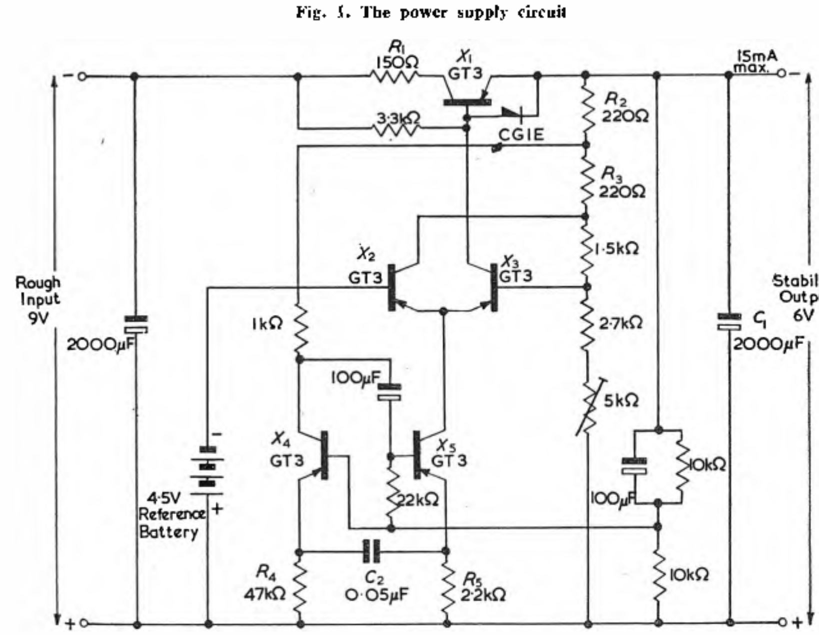
The circuit shown has a short term stability of about 10mV for any reasonable change of input voltage or load current. This arrangement, however, is intended primarily to illustrate the principles from which other designs could be developed. Almost any commercial type of junction transistor could be used in place of the B.T.H. types shown.

Yours faithfully,

K. C. Johnson,

Computer Department,
Ferranti Ltd.,
Manchester.

Fig. 1. The power supply circuit



BOOK REVIEWS

Frequency Modulation Engineering

By C. E. Tibbs and G. G. Johnstone. 435 pp. 255 figs. Demy 8vo. 2nd Edition. Chapman & Hall Ltd. 1956. Price 45s.

BOTH the authors have obviously been connected with sound broadcasting, and the subject matter of this book is almost entirely concerned with this aspect of frequency modulation. A less general title might have been found for a book which dismisses f.m., radio links, f.s.k. and picture transmission in ten pages of the final chapter, which could in expanded form make a good opening chapter for the book.

A minimum of mathematics is used and for this reason the book should have a wide appeal. All aspects of f.m. broadcasting are considered, some of them in more detail than is strictly necessary for the immediate purpose. After chapters on modulation, noise and interference suppression, more than a hundred pages are devoted to propagation and aerials. These chapters contain a useful review of the subject but much that has no special reference to frequency modulation. The second half of the book is devoted to transmitters and receivers for f.m. broadcasting. The importance of modulation circuits is stressed and methods of achieving the required frequency are described. A section is devoted to descriptions of several commercial transmitters, including a low-power single-channel equipment for link use. A separate chapter is devoted to limiters and discriminators followed by one on receivers; here again much useful material is included which is not particularly applicable to the subject. The section on intermediate amplifiers could have given more space to the question of phase distribution and less to the now familiar problems of Q and coupling coefficients. Special measuring techniques applicable to f.m. are briefly described including data on panoramic display methods.

The book is well printed with good illustrations, except for several drawings which are reproduced from other publications. Each chapter concludes with a list of appropriate references; the space is well used. More care could have been taken with the general revision of the book, particularly with the simple algebra, in which several mistakes were spotted. In the chapter on propagation a variety of units is used and in some cases the units are not stated; equations 5.9 and 5.10 are difficult to reconcile with each other and with previous work, whatever units are chosen. On page 111 a calculation of selective fading makes a substantial arithmetical error and then proceeds to confuse wavelength with frequency. It is not surprising that the result is wrong by a factor of 2. On the whole BSI abbreviations have been adhered to, but the HEN has been introduced in Fig. 46 as a unit of inductance.

G. L. GRISDALE.

The Principles of Colour Television

By The Hazeltine Laboratory Staff. Edited by K. McIlwain, C. E. Dean. 595 pp. 60 figs. Demy 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd., London. 1956. Price 104s.

THIS is a comprehensive exposition of colour television principles and, overall, the treatment is typical of the thoroughness usually associated with the Hazeltine Laboratories. The twelve authors, two of whom have edited this volume, have all made contributions to the new technique of colour television and it was, therefore, with some pleasure in anticipation that a reading of this book was undertaken.

The early chapters introduce the fundamentals of light and photometry, colour perception, colour space, colour triangles, and colorimetry. Most of this section of the book deals with the subject matter in a conventional way and underlines the general thought governing the choice of primaries for the N.T.S.C. system and clearly explains the *RGB* and *XYZ* co-ordinates systems and finally shows how these general principles are applied to the requirements of colour television.

The next chapters discuss the economies possible in transmission of colour information due to limitations and peculiarities of visual colour perception. The pre-correction of 'gamma' at the transmitter producing curves rather than straight lines on the chromaticity diagram could, perhaps, have been explained more fully at this stage, but in this case, as well as in most other cases, adequate references are given in a most comprehensive bibliography.

A number of chapters deal with the application of the general principles to the general engineering of a system.

Chapters on the production of the composite colour signal and synchronization deal thoroughly with these topics. Gamma correction and the implications of non-linear amplitude relations are fully discussed and a number of possible gamma correction systems are compared with that used on present 'N.T.S.C.' practice.

A statement and discussion of the N.T.S.C. colour television standards draws attention to the implication on equipment design.

Chapter XIII (51 pages) deals with equipment for producing the transmitted signal, including waveform generators, studio cameras, slide scanners and film cameras, and signal processing and coding equipment. Transmitters are disposed of rather arbitrarily by statements such as "any good monochrome transmitter can be adapted to colour operation by limited modifications and careful adjustment"—the whole treatment taking little over one page in a book of over 500 pages. This gives little guidance to the transmitter designer in respect of the additional VA required in the transmission system, the achieving of minimum differential phase error, etc.

In complete contrast the following chapters on colour receiver topics, extending to over 130 pages, are much more comprehensive; "decoders" are dealt with in 48 pages for 3-Gun Displays and a further 44 pages are devoted to decoders for One-Gun Displays (as against 5 pages for "coders" in Chapter XIII.)

The chapter on test and measuring methods is adequate and reasonably comprehensive. At the end of each chapter there is an impressive bibliography—these lists providing a total of over 500 references.

The book concludes with a glossary of colour television terms; two mathematical appendices and a third quoting the FCC Compatible Colour Order of December 1953.

Despite the somewhat unbalanced treatment of subject matter to which previous reference has been made, the book represents a mine of information which can be recommended without hesitation to all advanced students and practising engineers engaged on this most up to date branch of the television art.

V. J. COOPER.

Introduction to Colour TV

By M. Kaufman & H. E. Thomas. 154 pp. 60 figs. Demy 8vo. 2nd Edition. J. Rider Inc., New York, Chapman & Hall, London. 1956. Price 25s.

THIS introduction to colour television gives the newcomer to this new technique a simplified description of fundamentals, the N.T.S.C. system and, more particularly, of receivers designed for this system.

The general treatment of the subject matter makes this book of value to the user rather than the designer and should be particularly valuable to the service engineer.

The fundamentals are dealt with clearly and in sufficient detail for less technical readers and the authors have wisely avoided the complications of three dimensional colour space considerations and any of the usual "colour triangle" explanations. The treatment of fundamentals does, however, give the practical reader a good mental picture of the processes involved, including the differences between additive and subtractive colours in the general colour matching problem.

The broad principles of coding and decoding according to N.T.S.C. methods is clearly explained—the differences between 'I' and 'Q' working and '$B - Y$' and '$R - Y$' working being illustrated by simple vector methods. The description of colour tubes, both from a constructional and operational point of view, is well illustrated and covers both the 'shadow mask' type of tube and the Lawrence tube.

Receiver practices are given the greatest attention and the treatment is well illustrated with functional and detailed circuit arrangements, including a complete circuit diagram of the RCA CTC4 and C.B.S. 20S receivers, both of which utilize the RCA shadow mask tube. One, however, illustrates the full bandwidth I and Q demodulation system while the other illustrates the simplification that can result by using $B - Y$ and $R - Y$ demodulators at the sacrifice of some performance. Both these methods of operation are described clearly in the text.

Without making any pretext of being a book for the expert, this small volume gives practical introduction to present day systems and techniques and should be well received by those who find the more comprehensive volumes on this subject a little frightening.

V. J. COOPER.

Science and Information Theory

By Leon Brillouin. 320 pp. 60 figs. Demy 8vo. Academic Press Inc., New York. 1956. Price $6·80. Academic Books Ltd., London. 1956. Price 54s. 6d.

THIS book was written in the first place for physicists, but it must be warmly recommended also to communication engineers for whom information theory is more than just a modern formulation of 'signal-to-noise'. Professor Léon Brillouin is an outstanding physicist, who for more than 20 years is a leading personality in the development of shortwave and microwave communications, and has a full understanding of the difficulties of electrical engineers who want to master this by no means simple subject. Nothing is assumed about the reader except familiarity with the calculus and with the elements of probability theory. Everything else, from Fourier Series to Thermodynamics is derived and explained from first principles. The book is an almost complete compendium of the most important papers on information theory, even of its ramifications into physics, linguistics and semantics, and in almost every case the author has achieved remarkable feats of interpretation and of logical, yet easy presentation.

About half of the book is given up to the elucidation of the fundamental but by no means simple connexion between Shannon's measure of information and thermodynamical entropy, with a truly fascinating "exorcism" of Maxwell's Demons, which is for the most part the author's own work. The imprint of his hand is on every page, but particularly so in the chapters on Fourier analysis, including crystal analysis by X-rays, and on Brownian motion and noise. The brilliant discussion of noise in electric circuits may be recommended even to those engineers who do not wish to give up much time to communication theory, but would like to have a grasp of noise problems beyond the mechanical application of formulae.

The titles of some of the main chapters will convey an idea of the wide scope of the book:— Definition of Information. Redundancy in the English Language. Coding and Channel Capacity. Error Detecting Codes. Signal Analysis. Thermodynamics. Thermal Agitation and Brownian Motion. Thermal Noise in Electric Circuits. The Negentropy Principle. Maxwell's Demon. Writing, Printing, Reading, Computing.

It may be hoped that apart from introducing many engineers into this new and important field, this book will have a beneficial effect by correcting widely current misconceptions, due to the fact that there are so few electrical engineers who understand thermodynamics. The price is moderate, and students who buy this book will in all probability consult it again and again in their lifetime.

D. GABOR.

Television Servicing

By M. Mandl. 460 pp. 314 figs. Demy 8vo. Revised Edition 1956. Macmillan & Co, New York and London. Price 45s 6d.

THE author of this book is Director of Electronics and Television Courses at Temple University's Technical Institute. Each subject has been reexamined in the light of recent developments and brought up to date. These include the fundamentals of monochrome and colour television with diagrams of basic circuits, servicing instructions for u.f.h. and v.h.f., a section on series filament circuits, and an index to common television troubles which has been extended to include new references.

Radio, Volume 3

By J. D. Tucker and D. F. Wilkinson. 249 pp. 80 figs. Demy 8vo. English Universities Press. 1956. Price 12s. 6d.

THIS volume has been designed mainly to cover the requirements of the syllabus of the City and Guilds Radio 3 Examination. As in previous volumes, the subject matter has been broadened to include the relevant parts of such examinations as the Higher National Certificate (A.I.), the Advanced Radio Engineering, Radio Transmission and Radio Reception syllabuses of the British Institution of Radio Engineers and of the various examinations conducted by the Institution of Electrical Engineers.

Kempe's Engineers' Year Book 1957

3 000 pp. Crown 8vo. (Two volumes in case). 62nd Edition. Morgan Bros. (Publishers) Ltd. Price 82s. 6d.

THE 79 chapters of these volumes cover practically every branch of engineering. Most chapters have a short bibliography detailing the standard works on the subject. Useful additions have been made to various chapters and all have been revised where necessary.

ELECTRONIC EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

BEAT FREQUENCY OSCILLATOR

Distributed by B & K Laboratories Ltd, 57 Union Street, London, S.E.1

THE Brüel and Kjaer low frequency b.f.o. type 1015 has two frequency ranges, 2c/s to 2kc/s and 2002c/s to 4kc/s. Frequency accuracy is 1 per cent ±0·1c/s.

The frequency ranges, which are displayed on a logarithmic scale, can be swept automatically at a predetermined rate, by coupling the variable tuning capacitor via a flexible drive to a level recorder or other suitable motor. The drive is readily engaged and released by an electromagnetic clutch.

Provision is made to mark individual frequencies and to short-circuit the output during the return sweep of the capacitor.

The automatic output regulator enables the output voltage or current to be held constant. The instrument is well suited to provide the source for vibration generators, although a power amplifier will invariably be required between the 1015 and the vibrator. The output from a pick-up attached to the vibrator may be used as the control voltage for the automatic regulator, the acceleration, velocity or displacement thus being held constant. The time-constant of the control circuit may be adjusted to ensure operational stability.

NEW VALVES

(Illustrated above right)

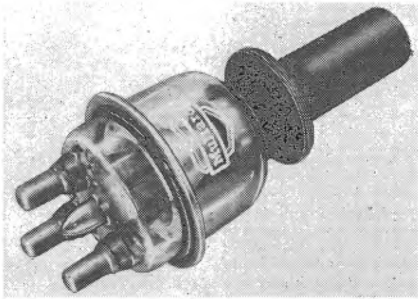
Mullard Ltd, Century House, Shaftesbury Avenue, London, W.C.2

THE QQV5-P10 is a double tetrode valve for use in the pulse modulators of radar and radio equipment. It is capable of delivering pulses of 5kV, 10A when both sections are connected in parallel. The valve is suitable for pulse durations of up to 3μsec and duty factors of ·001. The anode dissipation is 15W maximum and under typical conditions the total power input to the stage is 85W.

Type TY7-6 000, (illustrated) specially developed for use in industrial r.f. generators will deliver an output of 6kW at frequencies up to 50Mc/s and has been conservatively rated to ensure adequate margins of safety when used in industrial heating applications, where the valve may be subjected to mains and load variations, causing intermittent overloads.

The triode is available in two versions, TY7-6 000A which is designed for forced air cooling, and TY7-6 000W which is a watercooled version. The maximum permissible anode dissipation is 6kW in either case and anode voltages up to 7kV may be used. The valves have directly heated thoriated tungsten filaments and external anodes.

Valve QQVO2-6 will function efficiently at frequencies up to 500Mc/s, giving enough power output for mobile working.



The valve is a miniature double tetrode power amplifier valve on a noval (B9A) base and its centre-tapped heater can be operated from either a 12·6V, 0·4A or a 6·3V, 0·8A supply. A mutual conductance of 7mA/V per section has been obtained by incorporating a control grid having very fine grid wires positioned very close to the cathode. The grid-anode capacitances are internally neutralized, so that stable operation is obtained when the two sections are used in push-pull amplifier circuits.

As a power amplifier at 490Mc/s, the QQVO2-6 delivers 3·5W to the aerial under typical circuit conditions when operated at maximum ratings with anode and screen modulation. The drive power needed is 1·4W, and this can readily be obtained from another QQVO2-6 operating as a frequency tripler from about 163Mc/s. The h.t. voltage needed is 180V No special cooling arrangements are necessary.

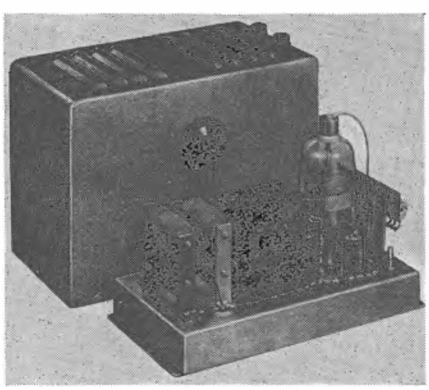
ELECTRONIC MOTOR CONTROL

(Illustrated below)

S. Brod & Co. Ltd, 31/33 Priory Park Road, London, N.W.6

DESIGNED for use with the Hillman F.H.6 1/4 h.p. motor, this electronic control has the following applications.

(a) The motor speed can be smoothly and continuously varied from 0 to 10 000rev/min maximum and is maintained within ±1 per cent of the setting for mains voltage variations of −10 per cent to +5 per cent (or wider variations of short duration), frequency variations of 47 to 65c/s and within ±3 per cent from off-load to full load.



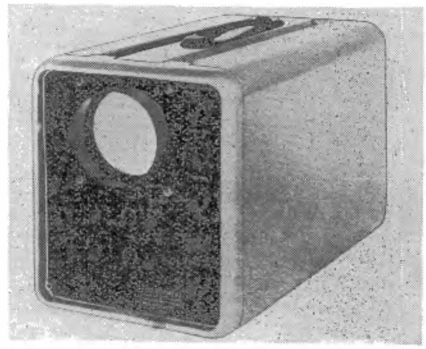
(b) By the addition of a tacho-generator, the speed is maintained to ±0·5 per cent or better for all variations.

(c) The speed of the motor can be controlled to maintain constant tension, pressure, temperature, thickness, light intensity, irradiation, etc.

(d) For coil winding and similar applications, a control which when switched on, will smoothly accelerate to a pre-set maximum speed and when switched off will smoothly decelerate. All three variables, i.e. acceleration, maximum speed and retardation are controlled independently.

Control is obtained by a xenon filled thyraton. The unit will operate satisfactorily at ambient temperatures of −20°C to +45°C with a delay of only 1min between switching on and full operation.

Control units can be supplied also for ½, 1 and 2 h.p. motors.



LABORATORY OSCILLOSCOPE

(Illustrated above)

Telequipment Ltd, 313 Chase Road, Southgate, London, N.14

TYPE 620 is a versatile oscilloscope with an 'Emitron' 4EP1 tube operated at 2·8kV and at the optimum p.d.a.-to-gun ratio.

The Y-amplifier is of the a.c. coupled constant-bandwidth type. The input is fed via a frequency compensated attenuator to a cathode-follower stage to achieve an input impedance of 2MΩ. The directly coupled push-pull output stage ensures a positive d.c. shift and freedom from distortion at full screen deflexion up to the highest frequencies. The sensitivity of 50mV per division is more than adequate for all wideband applications.

A novel form of voltage calibrator provides an accurate 1V peak-to-peak square wave enabling the Y-amplifier to be standardized, achieving an overall voltage measurement accuracy of ±5 per cent.

The time-base is a transitron oscillator which, although primarily intended for repetitive operation, has an adequate performance as a triggered time-base. The X-amplifier provides a minimum of five screen diameters magnification at all frequencies. A useful feature is the electrical interconnexion between the X-magnifier and the X-shift which ensures that the X-expansion is symmetrical about the centre of the screen rather than the centre of the trace, and the range of the X-shift is limited to the length of the trace at all settings of the magnifier.

CRANKSHAFT HARDENING PLANT

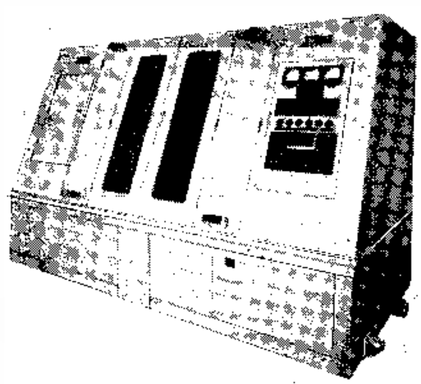
(Illustrated below)

Radio Heaters Ltd, Eastheath Avenue, Wokingham, Berkshire

THIS 'RADYNE' equipment is claimed to be the simplest and most economical method of processing a few hundred crankshafts per week. The plant is normally powered by the 'Radyne' C500 50kW output induction heating equipment which operates at a frequency of 450kc/s.

A heavy steel tank accommodates a jig plate on which are mounted the heating inductors for the particular type of crankshaft being hardened. Different crankshafts are handled by replacement of the complete jig plate and the drawer containing the process timers which fix the heating and quenching cycles for each particular crankshaft.

A control panel situated on the front



of the unit provides the facilities for automatic or manual operation. Once the crankshafts have been loaded, the heating and quenching cycle is completely automatic. The inductor coils are hinged and connected in parallel. Air rams close each coil in turn so that r.f. power is fed to only one coil at a time.

An automatic switching mechanism enables the heating and quenching cycles to be conducted on different bearings simultaneously. In addition, an automatic re-matching unit enables maximum loading to be maintained on the generator throughout the entire heating cycle, thereby running the plant at the maximum of efficiency.

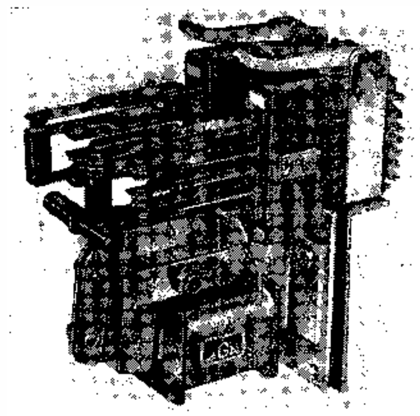
At the end of each heating cycle, water automatically floods over the heated surface and at the same time the r.f. power is switched to a coil on the other crankshaft.

PLUG-IN RELAYS

(Illustrated above right)

Ega Electric Ltd, Holyhead Road, Wednesbury, Staffordshire

THESE relays are built on the unit principle and all parts are easily replaceable, including the contact fingers.



The unit consists of a relay and terminal base, which can be mounted and wired into position before the relay proper is plugged-in, there being no soldered connexions to the relay itself.

The plug-in contacts are compressed in series and locked into position so that there is no likelihood of a faulty or light connexion, due to vibration or other causes.

The magnet coil supply leads are connected to the bottom plate of the finger boxes.

In this form, the relay occupies little panel space and avoids any chance of damage to the relay contacts during the wiring process.

The fingers and insulation pieces are assembled in a moulded box frame and compressed by set-screws against a squash plate, giving good contact pressure, and it is a simple matter to replace a bent or faulty finger by loosening the set-screws at the top of the box.

The whole finger box is fixed to the base by 2 screws accessible from the front.

Each standard finger box can accommodate up to 7 fingers, 6 for use as 3 pairs of normally open contacts, 3 pairs of normally closed contacts, or 2 sets of single pole change-over contacts, the 7th finger being used for the magnet coil supply.

Variations of the above up to a maximum of 6 fingers per box are possible.

It is possible to remove the complete magnet assembly after undoing two screws. The coil can then be removed once the main bearing pin is withdrawn and the coil retaining washer loosened.

The coils can be wound for all a.c. voltages from 4 to 440V and for use on d.c. from 6 to 230V.

STAFF LOCATION SYSTEM

Multitone Electric Co. Ltd, 12 Underwood Street, London, N.1

A SYSTEM has been developed in conjunction with St. Thomas' Hospital which operates by means of magnetic induction from audio-frequency currents fed by a transmitter into a loop round the building, and pocket receivers each of which responds to only one of 56 frequencies in the range transmitted.

The receivers, which clip into the pocket are about 5in (12·5cm) long × 1in (2·6cm) in diameter, weigh about 5oz (14·2g) with battery, and emit a note of about 1kc/s when the appropriate call comes through. They contain four transistors powered by a single Mallory cell. The quiescent current is less than 0·5mA, giving a cell life of over two months with the receiver permanently switched on.

Two versions of the receiver are available, one only gives the warning signal while the other, in addition, receives speech when a button is pressed. Each person on call is assigned a number of one, two or three digits, and to call him it is only necessary to press the buttons corresponding to those digits. The call terminates after one of the three periods which may be selected and all relays are released.

If it is desired to contact more than 56 persons several can share a single frequency and their calls be differentiated by coding.

In addition to the buttons for call selection, two coding buttons and a speech key are provided.

SIGNAL INTENSITY METER

(Illustrated below)

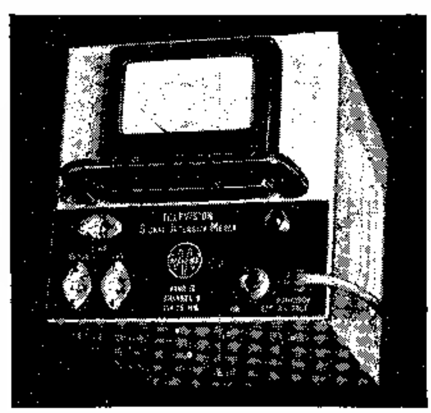
Radio-Aids Ltd, 29 Market Street, Watford, Hertfordshire

THE SIM3 is designed to give assistance in the erection, siting, orientation, comparison and fault finding in aerial systems.

The instrument is based on the use of straight r.f. amplification to provide freedom from second channel interference together with stability of calibration.

The instrument incorporates the principle of four cascodes in cascade, employing the Mullard PCC84 or its equivalent MAZDA 30L1 valves, followed by a CV425 germanium diode rectifier, thermistor-compensated for temperature variation, and a sensitive micro-ammeter. The latter is calibrated for direct indication of the amplitude of the signal applied from the aerial feeder to one of the three attenuator input sockets.

Full-scale sensitivity of the meter is set in the factory using an unmodulated carrier (equal to peak white amplitude) on the channel for the particular instrument, at 200μV, 2mV and 20mV respectively. The three ranges are selected by plugging an 80Ω concentric feeder into one of the three sockets provided on the attenuator. The calibration remains within 2dB over extended periods, the intensity scale is linear above 10μV.



Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 27 February. Time 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: Some Applications of Nucleonics in Medicine.
By: E. W. Pulsford.

North Western Section

Date: 7 February. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester, 1.
Lecture: Electronics in Medicine.
By: R. F. Farr.

West Midlands Section

Date: 13 February. Time: 6 p.m.
Held at: Wolverhampton and Staffordshire Technical College, Wulfruna Street, Wolverhampton.
Lecture: An Automatic System for Electronic Component Assembly.
By: K. M. McKee.

Merseyside Section

Date: 14 February. Time: 7 p.m.
Held at: The Council Room, Chambers of Commerce, 1 Old Hall Street, Liverpool, 3
Lecture: Radioactivity and its Measurement.
By: E. W. Pulsford.

Scottish Section

Date: 14 February. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: The Earth Satellite Project.
By: P. H. Tanner.
Date: 22 February. Time: 7 p.m.
Held at: Department of Natural Philosophy, University of Edinburgh.
Lecture: The Field Evaluation Trials of Electronic Equipment.
By: H. Holmes.
and: The Electronic Manipulation of Digits applied to Statistics.
By: J. Kyles.

South Wales Section

Date: 20 February. Time: 6.30 p.m.
Held at: Cardiff College of Technology and Commerce, Cathays Park, Cardiff.
Lecture: Radioactivity and its Measurement.
By: E. W. Pulsford.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Joint Meetings

Date: 7 February.
Third Graham Clark Lecture: The Place of Engineering in University Education.
By: Sir Ifor Evans.
Date: 25 February.
Held at: The Institution of Civil Engineers, Great George Street, Westminster, London, S.W.1.
Discussion on the Report on Engineering Education in the Soviet Union. (Copies of the Report can be obtained from the Institution.)

Radio and Telecommunication Section

Date: 4 February.
Informal evening on The Importance of Research in Hearing and Seeing the Future of Telecommunication Engineering.
Talk by: E. C. Cherry.
Date: 20 February.
Held at: Northampton Polytechnic, St. John Street, London, E.C.1.
Lecture: The Stereosonic Recording and Reproducing System. (A Two-Channel System for Domestic Tape Records.)
By: H. A. M. Clark, G. F. Dutton and P. B. Vanderlyn.

Faraday Lecture

Date: 13 February. Time: 6 p.m.
Held at: Central Hall, Westminster, London, S.W.1.
Lecture: Nuclear Energy in the Service of Man.
By: T. E. Allibone.

Measurement and Radio Sections

Date: 12 February.
Lectures: The Ultimate Performance of the Single-Trace High-Speed Oscillograph and The Design and Performance of a New Experimental Single Transient Oscillograph with Very High Writing Speed.
By: M. E. Haine and M. Jervis.

Informal Meeting

Date: 18 February.
Discussion: The Impact of Radioactive Isotopes on the Utilization Field
Opened by: Denis Taylor.

Measurement and Control Section

Date 26 February.
Discussion on The Analysis of Waveforms.
Opened by A. Cooper and D. A. Drew.

East Midland Centre

Date 6 February. Time: 7.15 p.m.
Held at: De Montfort Hall, Leicester.
Faraday Lecture: Nuclear Energy in the Service of Man.
By: T. E. Allibone.
Date: 12 February. Time: 6.30 p.m.
Held at: Loughborough College.
Lectures: Some Applications of Mercury Arc Rectifiers.
By: A. Wren.
and: The Transductor as Applied to Distance Protection.
By: H. Barber.

Cambridge Radio and Telecommunication Group

Date: 19 February. Time: 8 p.m.
Held at: The Cavendish Laboratory, Free School Lane, Cambridge.
Informal Evening on Electronics and Automation.
By: H. A. Thomas.

North-Eastern Centre

Date: 11 February. Time: 6.15 p.m.
Held at: The Neville Hall, Newcastle upon Tyne.
Supply Section Chairman's Address: Engineering Efficiency, Economics and Expediency.
By: P. J. Ryle.
Date: 18 February. Time 7 p.m.
Held at: The Carlisle Technical College.
Lecture: Automatic Circuit Reclosers.
By: G. F. Peirson, A. H. Pollard and N. Care.
Date: 19 February. Time: 7 p.m.
Held at: The Workington College of Further Education.
Lecture: Automatic Circuit Reclosers.
By: G. F. Peirson, A. H. Pollard and N. Care.

North-Eastern Radio and Measurements Group

Date: 4 February. Time: 6.15 p.m.
Held at: King's College, Newcastle upon Tyne.
Lecture: The Measurement of Earth-Loop Resistance.
By: G. F. Tagg.
Date: 18 February. (Time and place as above.)
Informal Lecture: The Use of Transistors in Radio and Television.
By: A. J. Biggs.

North Midland Centre

Date: 5 February. Time: 6.30 p.m.
Held at: Offices of the Central Electricity Authority, 1 Whitehall Road, Leeds.
Lecture: The Crystal Palace Television Transmitting Station.
By: F. C. McLean.
Date: 28 February. Time: 7.15 p.m.
Held at: Offices of the Yorkshire Electricity Board, Ferensway, Hull.
Lecture: Germanium and Silicon Power Rectifiers
By: T. H. Kinman, G. A. Carrick, R. G. Hibberd and A. J. Blundell.

Sheffield Sub-Centre

Date: 20 February. Time: 6.30 p.m.
Held at: The Grand Hotel, Sheffield.
Lecture: The Potentialities of Railway Electrification at the Standard Frequency.
By: E. L. E. Wheatcroft and H. H. C. Barton.

North-Western Radio and Telecommunication Group

Date: 13 February. Time: 6.45 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: Frequency-Modulated Quartz Oscillators for Broadcasting Equipment.
By: W. S. Mortley.

North-Western Supply Group

Date: 26 February. Time: 6.15 p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Lecture: Choice of Insulation and Surge Protection of Overhead Transmission Lines of 33kV and above.
By A. Morris Thomas and D. F. Oakeshott.

North Lancashire Sub-Centre

Date 13 February. Time: 7.15 p.m.
Held at The North Western Electricity Board Demonstration Theatre, North Road, Lancaster.
Lecture Germanium and Silicon Power Rectifiers.
By: T. H. Kinman, G. A. Carrick, R. G. Hibberd A. J. Blundell.

Northern Ireland Centre

Date: 12 February. Time: 6.30 p.m.
Held at: The Engineering Department, Queens University, Belfast.
Informal Evening on Electronics and Automation.
By: H. A. Thomas.

South-East Scotland Sub-Centre

Date: 5 February. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lecture: The Control and Instrumentation of a Nuclear Reactor.
By: A. B. Gillespie.

South-West Scotland Sub-Centre

Date: 6 February. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: The Control and Instrumentation of a Nuclear Reactor.
By: A. B. Gillespie.

South Midland Centre

Date: 4 February. Time: 6 p.m.
Held at: The College of Technology, Gosta Green, Birmingham.
Lecture: Technical Development and Training in the U.S.S.R.
By: G. S. C. Lucas and D. P. Sayers.

South Midland Radio and Telecommunication Group

Date: 25 February. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Recent Trends in Designs and Application of Closed Circuit Television Equipment.
By: I. M. Waters.

North Staffordshire Sub-Centre

Date: 11 February. Time: 7 p.m.
Held at: The Technical College, Stafford.
Chairman's Address.
By: A. T. Chadwick.
Date: 20 February. Time: 6.30 p.m.
Held at: The Electricity Service Centre, Burton-on-Trent.
Lecture: Automatic Circuit Reclosers.
By: G. F. Peirson, A. H. Pollard and N. Care.

Southern Centre

Date: 11 February. Time: 6.30 p.m.
Held at: The Guild Hall, Southampton.
Faraday Lecture: Nuclear Energy in the Service of Man.
By: T. E. Allibone.
Date: 20 February. Time: 7.30 p.m.
Held at: The R.A.E. Technical College, Farnborough.
Lecture: The Control of Nuclear Reactors.
By: R. J. Cox and J. Walker.

Western Utilization Group

Date: 25 February. Time: 6 p.m.
Held at: Electricity House, Colston Avenue, Bristol.
Lecture: Germanium and Silicon Power Rectifiers.
By: T. H. Kinman, G. A. Carrick, R. G. Hibberd and A. J. Blundell.

South-Western Sub-Centre

Date: 14 February. Time: 4 p.m.
Held at: Standard Telephones and Cables Ltd., Dowlish Ford Mills, Ilminster.
Lecture: The Generation and Synthesis of Music by Electrical Means.
By: A. Douglas.
Date: 21 February. Time: 3 p.m.
Held at: The Electricity Showrooms, Bedford Street, Exeter.
Lecture: Nuclear Energy and Electricity Generation.
By: G. T. Shepherd.

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