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## Commentary

AUTOMATION has been very much in the news in recent weeks. Discussion of this subject has, in fact, passed from the technologist to the layman. The whole subject has been grossly over-dramatized and many rash and erroneous statements have been made by persons who should be in a position to advise and lead their fellow workers. Certain organs of the national and provincial press have also failed to act with the air of responsibility which should form part of their duty to their readers. The subject has become cloaked in mystery and become the matter for grotesque and horrifying cartoons.

It is, of course, equally wrong to underrate the importance of automation or, indeed, to maintain that it is nothing new for, while labour saving and mechanical methods of one kind or another have been in existence since the beginning of history, automation as a total concept is certainly new.

If the recently published report\* which has been compiled by the Department of Scientific and Industrial Research, is sufficiently publicized and its contents assimilated by those concerned, it should do much to put automation in its true perspective. It is certainly of vital importance that both sides of industry should base their judgments on facts and here the D.S.I.R. report underlines one of the fundamental difficulties at the present time, and that is that little is really known of the consequences of the introduction of advanced automation; for at present this only exists to any great extent in the chemical and oil refining industries. In fact, the appreciation of the technical capabilities of automation has considerably outstripped our knowledge of the economic and social aspects. In a number of industries automatic control of individual processes is widespread and in some factories there is simple integration of control. Even now it is possible to visualize a number of these single processes joined together to form a completely automatic factory in which a computer integrates and controls the separate automatic production lines. To contemplate the economics of such an advanced state of automation is more difficult, while the sociological implications pose an even more difficult problem.

However, the report states "As experience of automation grows, its future importance to the nation becomes

increasingly apparent. Like other advances in technique, it will increase efficiency and should, therefore, reduce costs. It will be of special importance to a country so dependent on overseas trade as the United Kingdom, because it will increase production and help to keep prices competitive". If this be accepted as a correct statement, and it seems beyond argument that, if we are to compete in world markets we must use production methods at least as modern as our competitors, then the duty of the scientist and technologist is clear; every effort must be made to further the progress of automation and bring it to its logical conclusion.

That many problems, technical, economic and social-logical will arise is certain, but, as the report states, they are unlikely to prove catastrophic, providing that automation does not progress at too fast a speed. That the critical speed is not likely to be exceeded is guaranteed by at least one factor, the acute shortage of scientific and technical manpower, and this factor is common to every country in the world, including the U.S.A. and the U.S.S.R. Apart from the need for more technical manpower of all levels, the type of qualifications required may also undergo some change for the training of engineers will need to be broader as many of the technical problems encountered will call for a knowledge of various types of engineering such as mechanical, electrical and electronic. In addition, owing to the greater complexity of manufacturing techniques, scientists and technologists will be called upon to play a greater part in management than hitherto, and this again calls for a yet broader still education and training.

It would seem, from all the evidence, that automation must come and that the next decade or two will be a period of untold interest and adventure for those qualified to participate in this new phase of industrial enterprise.

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As from this month the price of ELECTRONIC ENGINEERING has been increased to 3s. It is with regret that the publishers have made this decision, but the cost of paper and printing, in common with all other commodities and services, have risen enormously in the last few years. A year's subscription will be 36s, the postage being paid by the publishers, so that the cost to a reader is the same whether the journal is ordered direct from the publishers or through a newsagent.

\* Automation (H.M.S.O., 1956).

# V.H.F. Communication by Ionospheric Scatter

By A. W. Cole\*, A.M.I.E.E.

*Propagation by means of ionospheric scatter is now being increasingly used for radio telegraph communication over distances up to 1 400 miles. A short review is made of the characteristics of propagation and typical transmitters, receivers and aerial systems are described. The article is based on work carried out by many authorities and a brief list of selected references is appended.*

FOR some twenty-five years it has been known that the signal strength received beyond the normal radio horizon from transmissions on frequencies above about 30Mc/s was frequently greater than could be accounted for by diffraction around the curved surface of the earth. Only during the past five years, however, has any serious attempt been made to utilize this phenomenon for communication purposes. It is now well known that a form of scattering of the radio waves takes place in two quite separate regions above the earth's surface, and in this way a small amount of energy is deflected downward which otherwise would be lost in outer space.

Scattering in the lower of these two regions—tropospheric scattering—is due to random variations in the refractive index of the air, and extends the possible range of transmission to some 500km although very great transmitter power would be necessary to ensure satisfactory reception at that distance. It is effective for frequencies up to at least 3 000Mc/s. The possible bandwidth of such transmissions is large enough for many simultaneous telephony channels and even sufficient for television; so that it is quite certain that there is an important future for this system for links where conventional line-of-sight transmission cannot be employed, and it is possible that when the technique is sufficiently developed it may be justified on economic grounds alone. Already tropospheric scatter links can be engineered with confidence, and their performance can be forecast with reasonable accuracy.

## Some Propagation and Systems Aspects

This article is concerned with transmission by the second type of scattering—ionospheric scattering. This is effective for frequencies in the lower part of the v.h.f. band, and is due to random variations in the ionization at a height roughly corresponding to that of the 'E' layer, which is well known in connexion with the transmission of radio waves at frequencies below 30Mc/s. While normal 'E' layer ionization is certainly due to solar radiation, there is evidence that other agencies, and in particular meteor trails, are responsible for the ionization which produces the scattered transmissions. The possible distance range is much the same as for 'E'-layer transmission of h.f. waves, say about 1 400 miles, taking into account refraction in the troposphere.

Although the received field-strength of an ionospheric scatter transmission varies continuously with time it is nevertheless possible to speak of a path attenuation referred to the long-term median field strength. The path attenuation increases sharply with increase in transmission frequency, so that frequencies above about 60Mc/s are

unlikely to be of use for this purpose. On the other hand, frequencies below 30Mc/s, and sometimes rather higher frequencies as well, can be transmitted by other ionospheric modes, so that much interference from other services is possible. Scatter transmissions are likely, therefore, to be limited to the band from perhaps 35Mc/s to 50Mc/s, with a preference to the lower frequencies in this band. The path attenuation does not vary greatly with distance between transmitter and receiver beyond about 600 miles where scatter transmission first becomes effective and up to the maximum distance noted above; this seems to be due to the effect of the increased path length being offset by the smaller scatter angle, i.e., the angle by which the wave has to be deflected downward at the scatter point.

For transmission distances less than the maximum, scattering may take place anywhere in a considerable region of the ionosphere "visible" from both terminals. The limits in azimuth of this region are set by rays having a total path length equal to the maximum scatter transmission distance; the limits in elevation are similarly defined by tangential rays from the opposite terminal. Since more than one scatter source will generally be active at any time there is a possibility of multiple signals being received with a maximum delay equal to the transmission time for the difference in distance between 1 400 miles and the shortest path distance between the terminals. For example, if the great-circle distance between the terminals is 1 000 miles, the maximum signal delay is about 2msec, which would represent a 40 per cent telegraph distortion on a 200-baud transmission. This multi-path transmission can to some extent be controlled by restricting the beam-width of the aerials, nevertheless, it sets a limit to the useful bandwidth which can be transmitted.

It seems, therefore, that while under favourable circumstances, limited availability telephony transmission may be possible by ionospheric-scatter transmission, the widest use for systems of this sort will be for telegraphic traffic. At least four teleprinter channels in time-division multiplex would usually be possible. The chief advantage over conventional h.f. ionospheric transmission is that, with suitable system design a sufficient signal strength is available at the receiver at all times, without diurnal or seasonal frequency change, and with comparative freedom from the influence of ionospheric disturbances.

In order to appreciate the receiving problem, it is necessary to consider in greater detail the nature of the received signal. Naturally, there is considerable variation on different circuits and at different times on the same circuit. Typically there is a background signal showing instantaneous fluctuations of the order of 10dB, with a superimposed diurnal variation of about another 10dB, and with frequent sudden increases in signal strength by as much as 40dB or more which may persist for several seconds. There are occasional deep fades

\* Marconi's Wireless Telegraph Co. Ltd.

due to cancellation of multiple signals. These characteristics are dealt with by designing the receiving equipment to accept a large range of input signals, and by using some form of diversity reception. Since the wave-front loses phase coherence over a few wavelengths transverse to the direction of propagation, space diversity with aerials some five to ten wavelengths apart in this direction is effective. A possible alternative to this is the use of wave-angle diversity, by directing the receiving aerials to different parts of the scattering region.

Because of the highly variable amplitude of the received signal, it is necessary to use frequency modulation at the transmitter. Since the sources of scatter ionization are often moving rapidly, a Doppler effect can occur giving rise to a received signal component at a frequency differing from the transmitted frequency by as much as two or three kilocycles per second. Although such Doppler-shifted components are usually of lower amplitude than those of transmitter-frequency, it is necessary to employ a frequency-shift for telegraph transmission greater than the Doppler-shift to avoid confusion; frequency shifts of 4kc/s or more are commonly employed.

It is readily possible to design a v.h.f. receiver such that, on a quiet site, the limit to the lowest useful signal-level is set by cosmic noise picked up by the receiver aerial. In order to ensure a minimum signal-strength sufficiently above this for distances approaching the maximum, it is necessary to use a transmitter-power of about 40kW, with a transmitting-aerial gain of about 20dB, corresponding to an effective radiated power of about 4MW.

#### Aerial Systems

An aerial gain of 20dB calls for an aerial aperture of about 17 square wavelengths, or some 10 000ft<sup>2</sup> at a frequency of 40Mc/s. There is, however, a limit to the effective increase in gain achievable by increasing the horizontal aperture and narrowing the horizontal beam-width, which can be explained on the assumption that there are a finite number of scatter sources active at any one time in the ionosphere. If the aerial beam is unduly narrowed, fewer sources are illuminated, and a lower field-strength is received. This is of particular importance during periods of low field-strength, when it has been found, for example, that increasing the beam-width and reducing the aerial gain may increase the field-strength substantially. On the other hand, if the beam-width is made too great, multi-path transmission becomes an operating limit. It is accordingly necessary to arrive at a compromise which takes into account the transmission distance and the required signalling speed.

As far as the aerial beam-width in elevation is concerned, there are practical difficulties in making this as narrow as could be desired. Aerials have been built with an aperture of only 3 wavelengths vertically, although at least twice this figure would be advantageous.

There has been a tendency to make receiving aerials identical to those used for transmitting, but since the bandwidth can be controlled by the beam-width of the transmitting aerials, it is permissible to use receiving aerials of greater beam-width. From the viewpoint of over-all system performance, the saving effected in this way by the use of cheaper receiving aerials can well be applied to the construction of additional aerials for triple or quadruple diversity.

In order to discriminate against interfering signals and noise, the receiving aerials must have radiation patterns as far as possible free from minor lobes. The same requirement for the transmitter aerials aims at the reduction of

interference to other services. Both transmitting and receiving aerials should have particularly good back-to-front ratios, in order to avoid the production or reception of delayed signals scattered forward from points behind the aerials.

Several different types of aerial systems have been used and proposed for ionospheric scatter transmission. For the maximum possible distances, the main beam of radiation must be as nearly as possible tangential to the earth's surface. This might appear to indicate that vertical polarization would be appropriate, since for horizontal polarization the electric field vanishes at a conducting surface. However, the field strength at low angles from vertically polarized aerials depends very much on the characteristics of the ground along the transmission path and serious losses in the neighbourhood of the aerials are possible; accordingly the performance of a horizontally-polarized aerial can be predicted with greater certainty. Furthermore, a vertically-polarized wave experiences a small extra loss in the scattering process as compared with a horizontally-polarized wave, so that, all things considered horizontally-polarized transmission is to be preferred.

Again, because of the very low angle of elevation of the radiation, the aerial sites have to be selected with careful consideration to the ground-reflected ray. The vertical phase-coherence of the wave-front is good enough to justify putting the aerial as near as practicable to the correct height which will align the ground-reflection lobe maximum on the scattering region. For a transmission distance approaching the maximum, this calls for a height to the mid-point of the aerial of 500 to 600ft. As it would usually be prohibitively expensive to elevate a high-gain aerial on masts or towers to this height above the ground, the choice of a site on the edge of a natural elevation or cliff is indicated. In the latter case, with the first Fresnel zone on a water surface, conditions would be near ideal, but considerable irregularities in the reflection surface can be tolerated. It is of equal importance that there should be no obstructions in the path of the beam visible from the reflection zone at elevations approaching that of the required ray, and this observation should take tropospheric refraction into account; this condition also would usually be met by a cliff-edge siting.

Rhombic aerials are generally unsuitable, mainly because of the difficulty in controlling side and back radiation lobes, but partly also because when designed for high gain the beam-width of the main lobe tends to be too narrow. In circumstances where they can be used, a low elevation for the main lobe can be realized with comparatively short masts by constructing the aerial on ground sloping downward in the transmission direction.

The type of aerial so far favoured has been a corner reflector with an aperture five wavelengths horizontally and three wavelengths vertically, and with a corner angle of 60°. The reflectors consist of multiple horizontal wires, while the radiator comprises four co-linear full-wave dipoles arranged along the axis, giving a horizontal beam-width of about 12°. The gain of an aerial of this type is about 19dB above that of a half-wave dipole in free space, and careful design of the reflectors enables a back-to-front ratio of better than -25dB to be obtained.

An alternative type of aerial which has been proposed consists of a broadside array of horizontal dipoles. This would have a horizontal aperture of four wavelengths and a vertical aperture of six wavelengths, thus approximately halving the free-space beam-width in the vertical direction, as compared to that of the corner aerial described above. The horizontal beam-width would be about 15°. With this

type of aerial a double reflector screen is necessary in order to achieve a front-to-back ratio of 25dB but the forward gain would be about 2dB greater than that of the corner aerial. Feeding the thirty-two folded half-wave elements forming the exciter curtain of this aerial would be a difficult but not insurmountable problem, while a point in its favour is that only two rather stout masts are required for its support, as against a total of nine masts for the corner aerial.

Yagi-aerial elements can also be combined in groups to give a gain of the order of 20dB. Such an array has only about half the number of feed-points as compared to the dipole array, and rigid tubular elements can be used instead of wires. The yagi-elements alone can be designed to give a good back-to-front ratio, and further protection in the backward direction can be provided by an additional aperiodic wire screen.

### Transmitters

The high power transmitters can be derived either from existing types of transmitters for the h.f. band or alternatively from transmitters designed for television using Band 1.

work or when there is a fault on one of the main transmitters.

The transmitters are driven from a basic frequency shift drive, the output of which is at 3.1Mc/s, which is already largely standardized practice for driving h.f. transmitters on frequency shift telegraphy or for single sideband telephony.

The output of this standard unit is applied to a mixer stage which is fed with the output of a crystal oscillator which determines the final carrier frequency. The mixer is followed by a frequency multiplier and amplifier which supplies 5W at the radiated frequency to a hybrid circuit which feeds the two independent transmitters or amplifiers.

The frequency shift drive carrier crystal oscillator and mixer and amplifier systems may be completely duplicated and an automatic changeover circuit provided so that in the event of failure of the drive, a spare drive is brought into service immediately.

The first stages of the transmitters contain phase delay controls, which enable the output phase to be set as required, according to the reading of a phase meter.

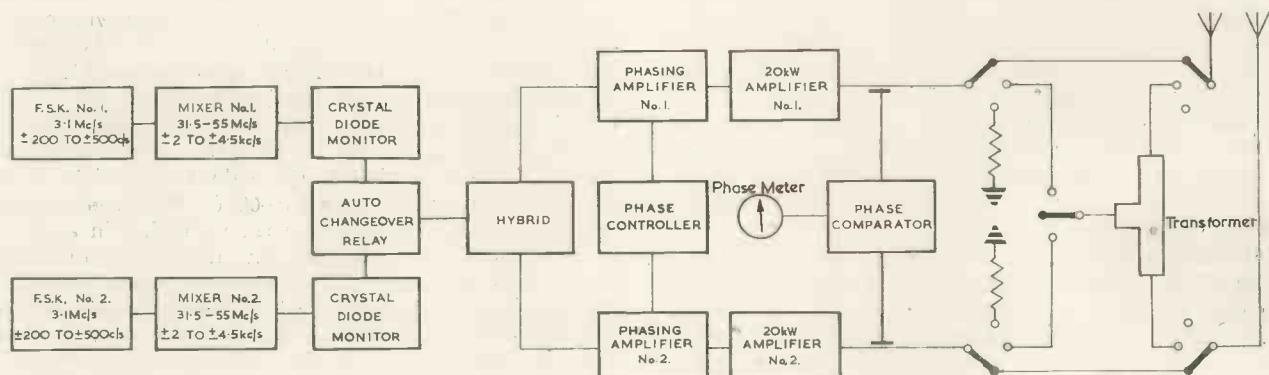


Fig. 1. Typical transmitter arrangement

A type of transmitter now being manufactured in the United Kingdom is of the latter type and is available in units having an output of 20kW. Where more than 20kW is required to establish a satisfactory circuit, two of these units are paralleled into a common aerial system in order to obtain 40kW output.

This combination of the two transmitter units may be effected directly on to one aerial array or alternatively the transmitters may each be connected to separate but identical halves of the array which are carefully erected so that no phase difference is introduced in the composite radiated signal. In either case to cater for random phase shift in the transmitters it is necessary to have some means of measuring phase difference at the output of the two transmitters and of adjusting it either to zero or to some predetermined value required for a given radiation pattern.

Fig. 1 illustrates the latter arrangement. The two halves of the array are erected in a broadside manner thus reducing mutual coupling and ensuring that the input impedance of either aerial is unaffected by the conditions in the other. This is important if it is desired to operate only one half of the array with one of the transmitters. In the event of only one 20kW transmitter being available means are provided for feeding the complete array.

In the case of very important channels it is possible to have a third transmitter available which can be brought into operation when it is required to carry out maintenance

Phase indication is provided by comparing pick-up from three points in the feeder system. Three voltage vectors of equal amplitude are derived, one at 0°, one at 180°, and a third at nominally 90°. The anti-phase pair are from one transmitter output feeder and the quadrature vector from the other transmitter output feeder. The deviation from the nominal 90° of the quadrature vector is a measure of the phase difference between the output signals and is obtained by arithmetical subtraction of the resultants of vector addition of 0° + 90° vectors and 180° + 90° vectors. Adjustment of the fine phasing control to produce zero indicates that the transmitters are in phase.

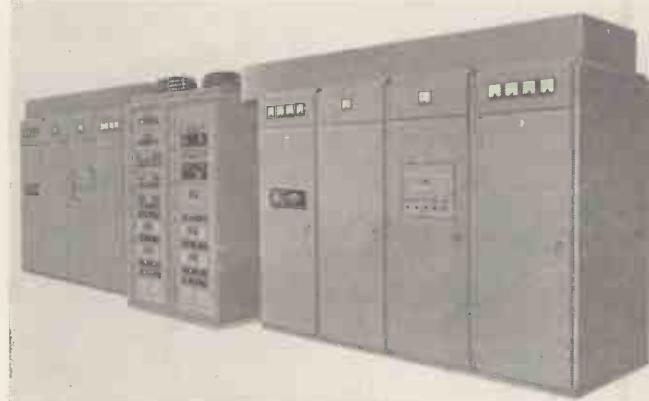
Phase can be measured up to 50° maximum and the adjustment provided gives co-phasing with an accuracy of better than 1°.

Each 20kW radio frequency amplifier is housed in four cabinets occupying a total floor area of 10ft by 3ft 9in, the h.t. transformer, smoothing circuits and blower being mounted externally. The radio frequency stages of the amplifier are housed in one cabinet 2ft 6in wide by 3ft 9in deep by 7ft high and comprise three stages of push-pull tetrodes valves which raise the power to 20kW. The first stage is a type QV0640 valve, double tetrode, which drives an intermediate amplifier of 4 valves type QY4-250 arranged in parallel push-pull. The final amplifier has two type CR192 radiator-cooled tetrodes, coupled to a 50Ω unbalanced feeder. All circuit elements are normal capacitor

and inductor type, transmission line technique being considered too elaborate for the particular service which these amplifiers had to meet. The resulting design has shown itself to be simple in construction, relatively low in cost and reliable in operation.

The h.t. rectifier consists of six mercury pool grid controlled tubes arranged in a three-phase half-wave circuit. Grid control is used only for the switching operations, not for close voltage control, and enables the rectifier to be run at one-third or two-thirds or full voltage as desired. The smoothing is inadequate under the first two conditions, but they are provided for use in tuning the amplifier only. For normal operation all tubes are excited. The overload protection circuits also operate on the grids of the rectifier tubes; because of this facility "on-load" star/delta switching contactors of high rating are not required. In fact, in each amplifier, apart from filament contactors, there is only one h.t. isolating contactor, which normally does not operate on power. Considerable field service has shown this particular type of rectifier to be most reliable and its control facilities to be most satisfactory.

Control facilities are such that the amplifiers may be either controlled remotely by a single push button or locally by interlocking push buttons enabling the amplifier to be run up to any required stage for checking and maintenance.



*Typical 40kW ionospheric scatter transmitter—Marconi type HS.201*

Overload indicator lamps immediately indicate to the maintenance staff the location of faults.

The equipment is air cooled, the circulation being by suction from the top of the transmitter, filtered air being admitted from a duct under the floor of the transmitter hall. The constructions of the cabinets are such that they are dust proof and the doors are all fitted with rubber sealing gaskets.

In its standard form the transmitter will operate over the frequency range of 31.5 to 55 Mc/s and will radiate a frequency shift telegraph signal with a shift of between 4kc/s and 9kc/s.

Normally only telegraph transmission is used, but if telephony communication should be required simultaneously then the radio frequency amplifiers are operated under class-B conditions and the frequency shift telegraph signals and the frequency modulated telephone signals are passed through the amplifier simultaneously. To effect this combination it is usual to provide duplicate mixer circuits fed from the common primary drive source, the telegraph modulation appearing on one side of the nominal carrier and the telephone modulation on the other.

Direct multiplexing of the transmitter on a straightforward

frequency modulation basis, with the telegraph signals accommodated in the band 0 to 300c/s and telephone transmission in the band 300 to 3 000c/s, is possible, but this would limit the telegraph transmission to one channel of 50 bauds only. Where time division telegraph systems have to be accommodated then the linear amplifier solution with two separate transmissions is to be preferred.

It is rare, however, that propagation conditions permit the transmission of telephony.

### Receivers

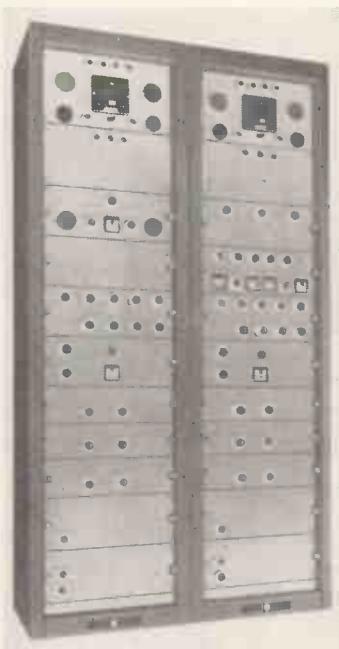
It is normal practice to transmit and receive from a common site if ionospheric scatter circuits are to be worked with reasonable efficiency, otherwise the telegraph receiving practice tends to follow that which has been established for systems in the high frequency band. As already mentioned, however, there are basic differences in the type of radio signal to be received. In the h.f. band seasonal and diurnal variations of signal level are very great as are also the signal-to-noise ratios and the short-term variations of signal level can be extremely erratic and severe.

On ionospheric scatter circuits the median level of the signal is much steadier and short-term fades of about 10dB represents the worst fading conditions. The signal-to-noise ratio, although poor, is reasonably constant. There are, however, large and sudden increases of field strength which may last for a few seconds up to as much as half a minute and which have already been commented upon and which are probably caused by meteor trails. These sudden increases in received level are often more than 40dB.

Such bursts are also associated with Doppler effects, and low level beat effects with the main carrier are common.

Normally, systems required for high speed frequency shift telegraphy, with a speed of 200 bauds, can be operated within a bandwidth of about 1kc/s. This pre-supposes the definite inclusion of the third harmonic of the telegraph waveform and a frequency shift of 500c/s or less. While these conditions can be attained on systems working in the high frequency bands, for ionospheric scatter reception Doppler effects make it essential to use a frequency shift at least in excess of 4kc/s. This is a drawback in that for telegraphy the bandwidth of the system is unnecessarily increased, and a new design approach is necessary if the performance of the system is not to be unduly affected.

A convenient solution to the problem is to discontinue the use of the conventional linear discriminator (which is usually designed to be linear over the full band of the receiver) and substitute a type of receiver which selects the mark and space signals only, largely ignoring the intervening region between them.



*Double diversity telegraph receiver for ionospheric scatter systems—Marconi type HR.16*

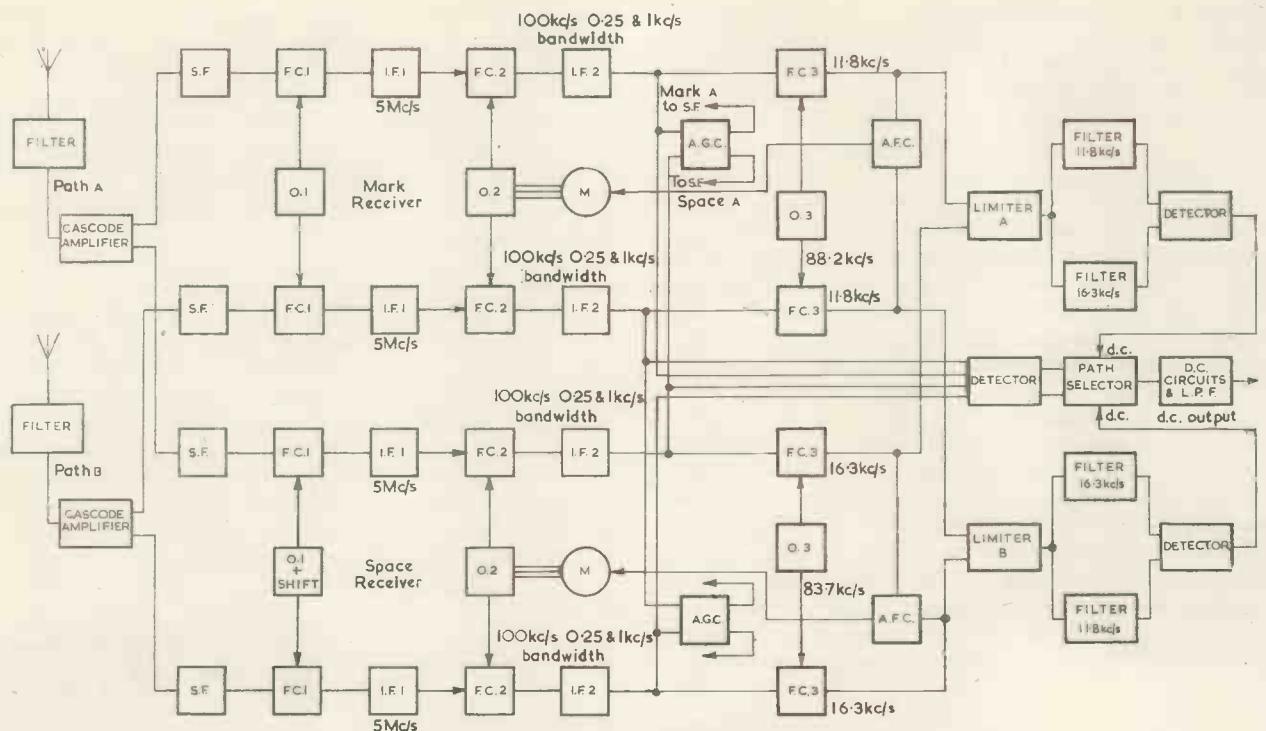


Fig. 2. Arrangement of double diversity receiver

Fortunately, this can readily be done using two receiver chains, each having the standard type of bandwidth as used in high frequency systems. As a matter of convenience and because of the advantages of flexibility and simplicity of operation, separate diversity receivers are therefore used for the mark and space frequencies, the outputs of these units being fed into a suitable path selector and combining circuit. The diversity path selector system operates on the principle of selecting the path having the superior signal-to-noise ratio, but in the event of the paths approaching equality then the amplitudes of the signals in the two paths are added.

A block schematic of a double diversity receiver is shown in Fig. 2. From the feeder system the incoming signal is fed through a high-pass filter, the purpose of which is to reject unwanted signals from the transmitter (which is usually on the same site) and other unwanted interference. This filter is adjusted to provide at least 60dB rejection to the transmitted frequency and a typical characteristic is shown in Fig. 3. The filters are followed by low noise cascode amplifier stages having a noise factor of approximately 4dB; these stages have duplicate outputs which feed the mark and space paths of the receivers respectively.

Quadruple detection receivers are used having a first intermediate frequency of 5Mc/s, a second intermediate frequency of 100kc/s, and final frequencies of 11.8kc/s for the mark receiver and 16.3kc/s for the space receiver; this arrangement being to facilitate diversity combination. The use of 100kc/s for the second intermediate frequency is common practice in h.f. telegraph receivers.

The first intermediate filters have a bandwidth of 25kc/s and the second intermediate frequency filters may have alternative bandwidths of 250c/s, suitable for a single printer channel, and 1kc/s which is suitable for a 4-channel printer system. The filters are designed for signalling speeds of 50 bauds and 200 bauds respectively and the fifth order sideband of the telegraph signal is only attenuated by 6dB.

Each receiver, i.e. the mark receiver and the space receiver, has its own common oscillators throughout, with

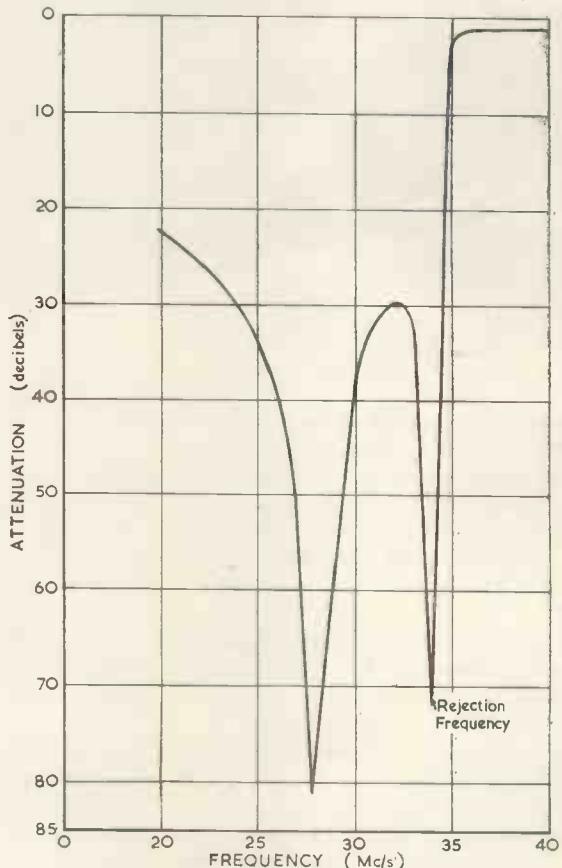
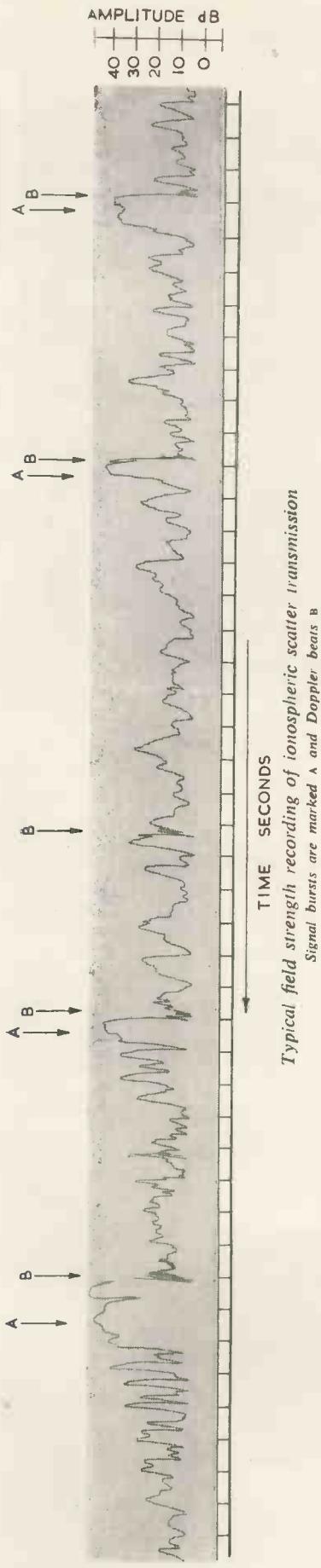


Fig. 3. Typical aerial filter characteristic

its own automatic frequency control system. This feature has advantages in that it provides for greater flexibility of operation, it being possible to change the shift to any reasonable figure without complicated changes at the



receivers. In the type of receiver where the mark and space signals are separated at low frequency, and where common r.f. and i.f. chains are used, provision for an operational change of frequency shift often leads to complications.

The maximum capture region for the a.f.c. system is 3kc/s and the accuracy of the output frequency into the limiter is better than 10c/s on each receiver.

After the third frequency changer, where the output of the mark receiver is 11.8kc/s and where the output of the space receiver is 16.3kc/s, paths A and B are reconstituted in a limiter for each path.

The limited signals from each path are then fed into two demodulating filters which drive a polar detector. The filters have a bandwidth of 500c/s for 200 baud operation and 125c/s for 50 baud operation. The path selector-combiner for each path accepts a direct current output from the two demodulators. After combination or selection and low-pass filtering the resultant telegraph signal controls a d.c. amplifier which can operate a telegraph relay, teleprinter or other telegraph apparatus.

Path selection or combination is determined by a comparator circuit which examines the output signal from the second intermediate frequency amplifier.

Over the primary working range of the path selector, the contribution to the output by either path is a function of the relative signal-to-noise ratios of the two paths. For signal levels outside this primary working range the path

TABLE 1.

SIGNAL CONDITION AT I.F.2 OUTPUT						OUTPUT CONTRIBUTION	
PATH A			PATH B			PATH A (PER CENT)	PATH B (PER CENT)
Signal (mV)	Noise (mV)	S/N (dB)	Signal (mV)	Noise (mV)	S/N (dB)		
37.5	12.5	10	37.5	12.5	10	50	50
"	"	"	37.5	18.75	6	63.5	37.5
"	"	"	37.5	37.5	0	100	0
"	"	"	25.0	12.5	6	72	28
"	"	"	12.5	12.5	0	100	0

selector operates in one of two ways. With very low levels the two signals are added and contribute equally to the output, while with high signal levels one path or the other is selected. An example of the path selector operation within the primary working range is given in Table 1.

### Conclusion

It will be realized from the foregoing that the terminal equipment for an ionospheric scatter circuit is likely to be more costly than that for a corresponding system working in the high frequency band and that the equipment is less flexible in use.

On the other hand the reliability of an ionospheric scatter system is likely to be greater than that of an h.f. system, although this advantage will vary considerably, depending on the location of the system.

The high cost of the transmitters and aerials, together with the lack of flexibility and difficulties in finding suitable sites may well make ionospheric scatter systems at their present stage of development unattractive for commercial users. For most users requiring point-to-point communication the high frequency band will probably continue to be more suitable, in that although the overall reliability may be less, a system working in this band is more economical to operate, requires less precise engineering planning, and permits of greater flexibility in the use of transmitting and receiving plant.

There will, however, be circumstances where operation in the high frequency band is difficult, due to the transmission path being in the auroral zone, or because of susceptibility to interference, or for other reasons; and then, for circuits of major importance, or for strategic communication, transmission by way of ionospheric scatter will be indicated.

### REFERENCES

- ECKERSLEY, T. L. An Investigation of Short Waves. *J. Instn. Elect. Engrs.* 67, 992 (1929).
- ECKERSLEY, T. L. Studies in Radio Transmission. *J. Instn. Elect. Engrs.* 71, 405 (1932).
- BOOKER, H. G., GORDON, W. E. A Theory of Radio Scattering in the Ionosphere. *Proc. Instn. Elect. Engrs.* 38, 401 (1950).
- MEGAW, E. C. Scattering of Electromagnetic Waves by Atmospheric Turbulence. *Nature, Lond.* 166, 1100 (1950).
- BAILEY, D. K., BATEMAN, R. et al. A New Kind of Radio Propagation at Very High Frequencies Observable over Long Distances. *Phys. Rev.* 86, 141 (1952).
- RATCLIFFE, J. A. Diffraction from the Ionosphere and the Fading of Radio Waves. *Nature, Lond.* 162, 9 (1948).
- ISTED, G. A. Atmospheric Electricity and Long-Distance Very High Frequency Scatter Transmissions. *Marconi Rev.* 17, 37 (1954).
- BRAY, W. J., HOPKINS, H. G., KITCHEN, F. A., SAXTON, J. A. Review of Long-Distance Radio-Wave Propagation above 30Mc/s. *Proc. Instn. Elect. Engrs.* 102, Pt. B, 87 (1955).
- BRAY, W. J., SAXTON, J. A., WHITE, R. W., LUSCOMBE, G. W. VHF Propagation by Ionospheric Scattering and its Application to Long-Distance Communication. *Proc. Instn. Elect. Engrs.* 103, Pt. B, 236 (1956).
- BAILEY, D. K., BATEMAN, R., KIRBY, R. C. Radio Transmission at VHF by Scattering and other Processes in the Lower Ionosphere. *Proc. Inst. Radio Engrs.* 43, 1181 (1955).
- ABEL, W. G., DE BETTENCOURT, J. T., CHISHOLM, J. H., ROCHE, J. F. Investigations of Scattering and Multipath Properties of Ionospheric Propagation at Radio Frequencies exceeding the MUF. *Proc. Inst. Radio Engrs.* 43, 1255 (1955).
- SUGAR, G. R. Some Fading Characteristics of Regular VHF Ionospheric Propagation. *Proc. Inst. Radio Engrs.* 43, 1432 (1955).

# A Portable Digital Frequency Meter

By M. I. Forsyth-Grant\*, A.M.I.E.E.

*A description of a portable frequency-time measuring instrument is given. It will operate at frequencies of up to 100kc/s showing the last four figures in discrete digits on an illuminated escutcheon. Overall accuracy is the order of five parts in  $10^5$ , this being derived from an internal crystal secondary standard. Hard valves are used throughout the instrument.*

RECENT advances in the radio, communication and industrial electronic fields have made the necessity for accurate measurement of frequency almost essential, and to meet this need a number of instruments of the digital counting type have appeared in most manufacturing countries. However, many of these are rather large and cannot be readily carried about the laboratory or workshop. The instrument about to be described is of a portable type, being about half the size and weight of a conventional oscilloscope. It has an accuracy better than five parts in  $10^5 \pm 1$  cycle, but it is possible to "standardize" the internal 100kc/s master crystal such that the accuracy is then of the order of one part in  $10^6$ . The overall weight is 32lb, and there are in all 37 valves; power consumption is approximately 120W. Fig. 1 shows a general view of the instrument.

## Principle of Operation

A block diagram of the digital frequency meter is shown in Fig. 2. Input pulses which may be sine or rectangular are amplified and then formed into uniform pulses of suitable shape to operate the counting chain through the signal gate. This gate is arranged to be opened for a period of one second under control of the time-base and control circuits. During the period of one second that the signal gate is opened, the counting chain totalizes the pulses passing through, and thus a reading of pulses per second is obtained.

## Design and Construction

The instrument may be sub-divided into the following main sub-assemblies:

- (1) The counting chain
- (2) The signal gate
- (3) The time-base
- (4) The control circuits
- (5) The power unit.

### COUNTING CHAIN

The counting chain consists of a number of decade count-

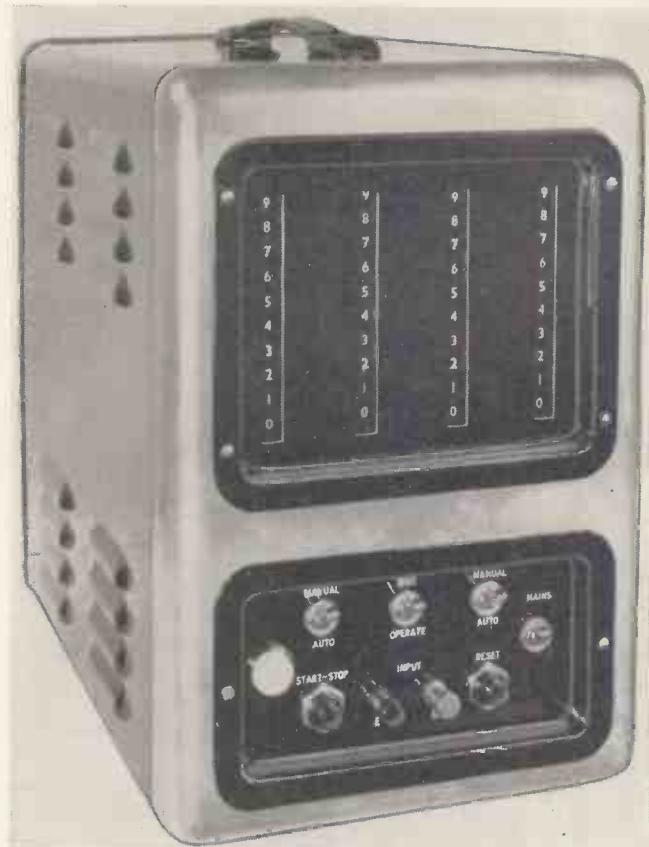
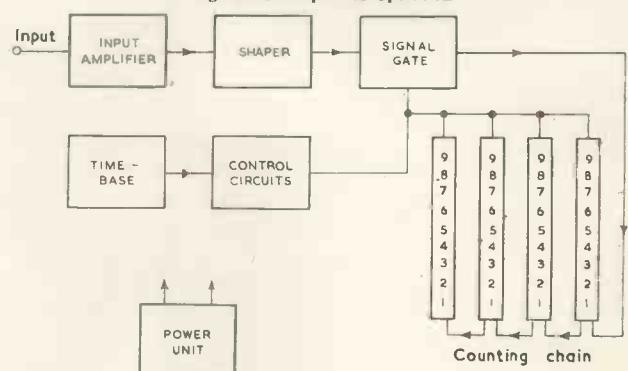


Fig. 1. The digital frequency meter

ing units and these make use of the well-known bistable multivibrator due to Eccles-Jordan<sup>1</sup>, commonly called a binary. A total of four decade counting units are used and each of these has four binaries in cascade, the valves being of the hard type, double triodes, throughout. A full description of the principles involved has been published elsewhere<sup>2,3,4,5,6</sup> so it will be sufficient to say that the decade counting units have the ability to count from 0 to 9 and then reset to 0 and at the same time to "carry one" to the next counting unit. Thus if a number of decade counting units are cascaded together a true decimal counting chain is established. Input in the form of electrical pulses is fed into the "units" counter unit at the right-hand side. At the count of 10 it transfers a pulse to the "tens" counter unit and so on. Each counting unit will operate at frequencies from 0 to 115kc/s but in order to conserve space and weight in this particular instrument, only four such counting units are provided, the first two significant

Fig. 2. Principles of operation



\* Racial Engineering Ltd.

figures of the frequency to be measured being usually known in advance.

A typical counting unit is shown in Fig. 3 in which the case has been removed to show the assembly which employs printed wiring throughout except for the leads to the valve holders and plug-base. By this means, great uniformity of production is achieved, this being particularly important to ensure that stray capacitances are kept to a minimum.

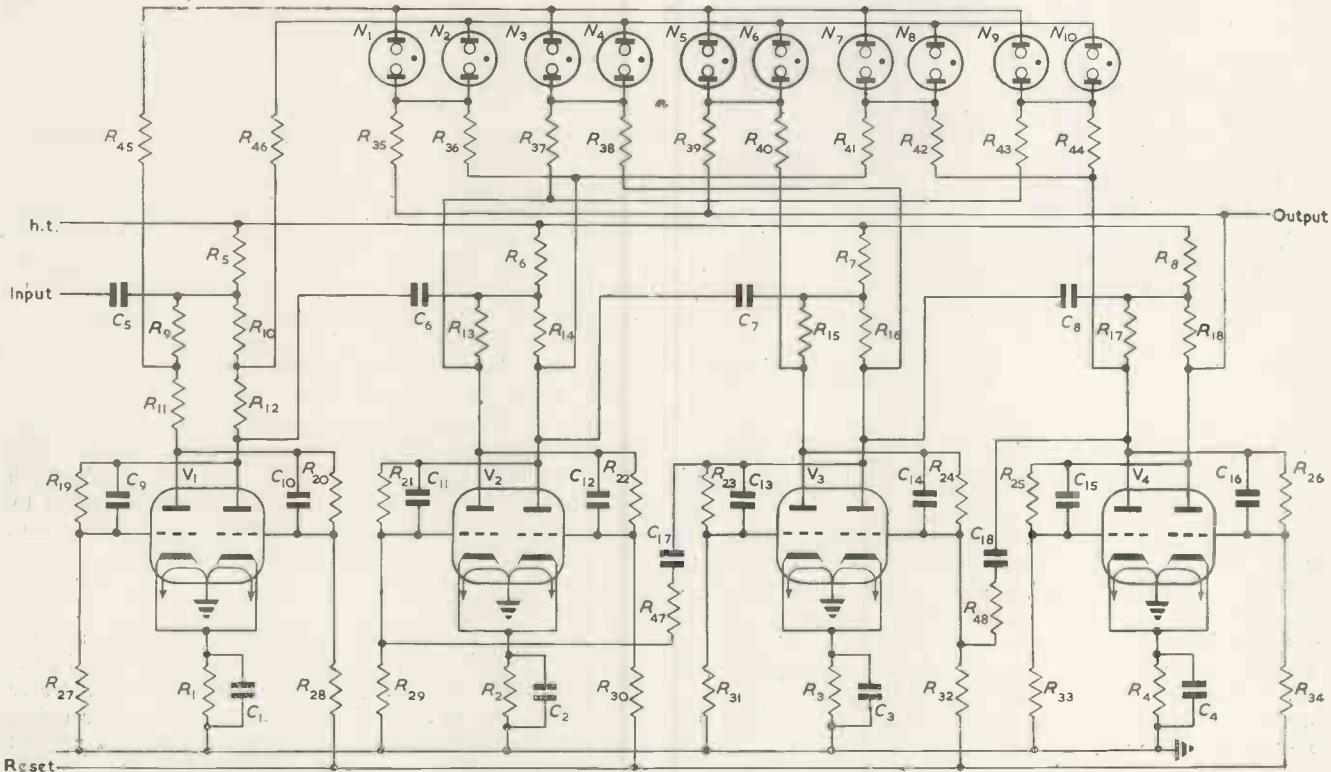
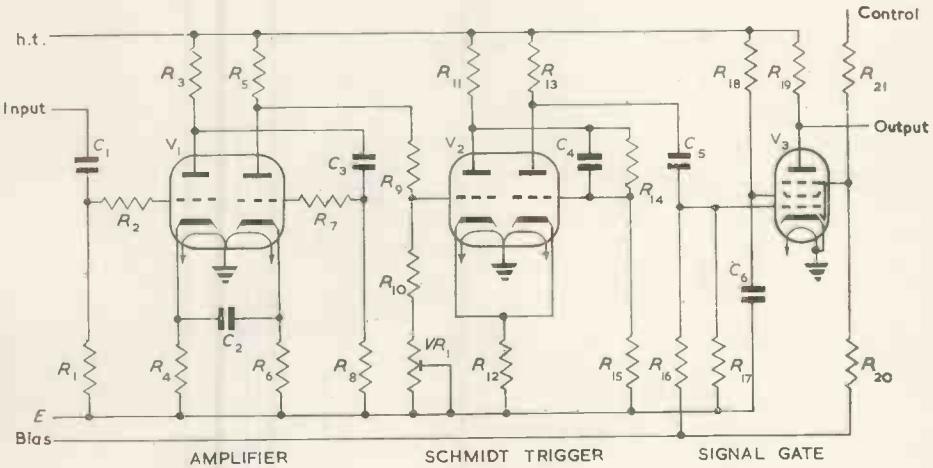


Fig. 3 (left). Digital counting unit showing printed wiring

Fig. 4 (above). Digital counting unit

Fig. 5 (below). Signal gate



A circuit of a typical counting unit is shown in Fig. 4. The natural scale of 16 of the unit is resolved to a scale of 10 by means of networks  $C_{17}R_{47}$  and  $C_{18}R_{48}$ . The first network is responsible for adding an additional "two" by changing over the binary stage  $V_2$ , and similarly, the other network, by operating binary stage  $V_3$ , adds "four" to the

count. Thus by adding six additional counts artificially, a scale of ten or decimal system is derived. Neon lamps  $N_1$ — $N_{10}$  give the progress of the count from 0-9. All units may be reset to 0 by momentarily opening the reset line from earth.

#### THE SIGNAL GATE

To convert a counter into a frequency meter, it is neces-

sary to control the time period during which signals are applied to the counting chain. It is the purpose of the gate to permit signals to pass to the counting chain, the gate itself being in turn controlled by the time-base and control circuits. The gate valve used in this instrument is a short base suppressor grid pentode which may be seen as  $V_3$  in

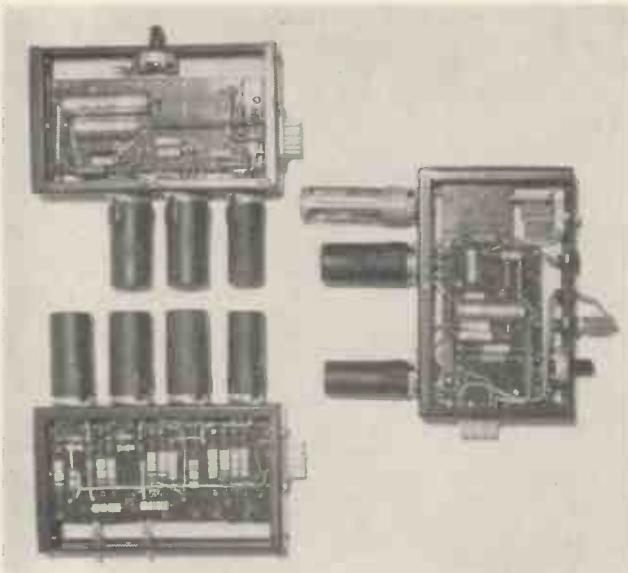


Fig. 6. Plug-In assemblies

Top left—Signal gate with its associated pre-amplifier and shaper. Right—Crystal oscillator and first phantastron divider. Bottom left—Control circuits

Fig. 5. The input signal, after being suitably amplified by  $V_1$ , shaped by Schmidt trigger  $V_2$ , is fed on to the first grid of  $V_3$ . This valve, Mazda type 6F33, has the useful property of being capable of being turned on and off by a small change in its suppressor grid potential. Thus the signal from the anode is obtained when the suppressor grid voltage is at about earth potential, but is absent when the suppressor is biased at  $-30V$ . This voltage change is obtained from the time-base, so that the gate is only opened for exactly one second. For the sake of convenience, the input amplifier  $V_1$ , the shaper  $V_2$ , and the gate valve  $V_3$ , are manufactured in a single plug-in unit. Fig. 6 shows the photograph of this.

#### TIME-BASE

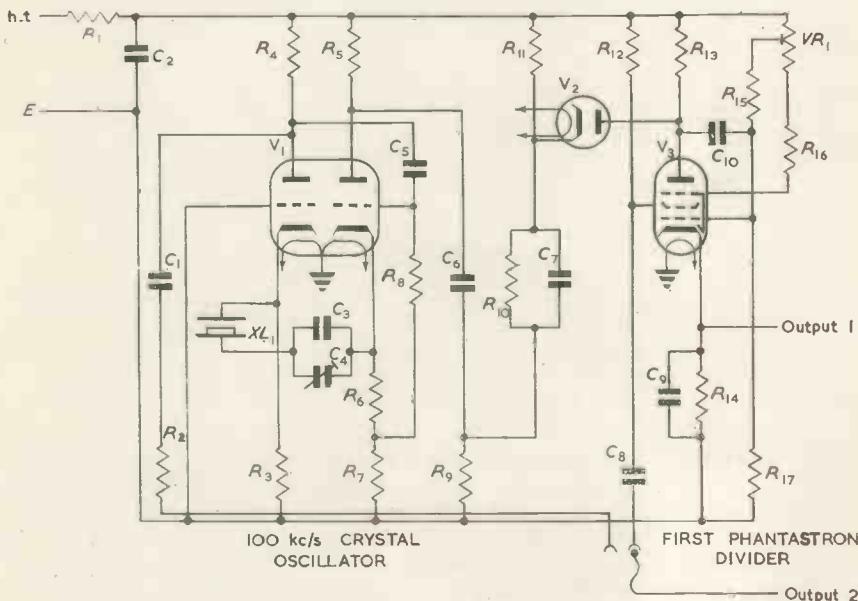
It is the purpose of the time-base to produce the time interval of one second during which the input signal is applied to the counting chain. The standard frequency used in this instrument is 100kc/s crystal, operated in a

series mode cathode coupled oscillator<sup>7</sup> ( $V_1$  in Fig. 7). The 100kc/s signal is differentiated by  $C_6$  and  $R_9$  and then fed into a Phantastron type divider<sup>8</sup> circuit  $V_2V_3$ , which provides an output pulse for every ten input pulses. There are in all five Phantastron dividers, each dividing by a factor of ten. The crystal oscillator and one Phantastron stage are mounted in one plug-in unit (Fig. 6), the remaining four Phantastron stages in another. The last Phantastron provides suitably shaped pulses—one per second. These Phantastrons are very stable in operation, even with the high division ratio of 10:1 due to their "linear" run down characteristic. Further they are very free from jitter. The time-constant of the run down is determined in  $V_3$  (Fig. 7) by capacitors  $C_{10}$ , resistor  $R_{15}$ , and preset potentiometer  $VR_1$ . The use of high stability capacitors and resistors, each with very small temperature co-efficient, is particularly essential in this instrument, for being constructed in portable form, the temperature rise is considerable. This might affect such a change in the  $RC$  time-constant as to make the divider ratio change to a factor of nine or 11 instead of the desired ten. However, the Phantastrons have a locking bandwidth of approximately  $\pm 3$ kc/s and therefore all time-base units are factory checked to ensure that the "locking" bandwidth is from 97kc/s to 103kc/s. These figures giving adequate safety margin for the instrument to operate indefinitely after approximately one minute warm-up period. In time, the capacitors and resistors associated with the Phantastron may change slightly but this may be rapidly corrected by adjustment of the preset potentiometer  $VR_1$  which allows for a margin of  $\pm 5$  per cent in circuit constants.

#### CONTROL CIRCUITS

A diagram of the control circuit is shown in Fig. 8. Before explaining this it is assumed that all circuits are reset and ready to start a frequency measurement. Valve  $V_1$  is known as the "time-base gate", which has the pulses from the time-base, spaced at one-second intervals, applied to its control grid.  $V_1$  is in a state of conduction in that the suppressor grid is at such potential as to permit signals to pass to its anode in a similar manner to the signal gate described above. The first pulse appearing at the anode of  $V_1$ , is fed through  $C_3$  to binary valve  $V_3$ , which immediately changes over to its other stable state. The left-hand anode of  $V_3$  goes positive and being connected to the signal gate suppressor grid ( $V_3$  in Fig. 5) via the right-hand cathode-follower stage  $V_2$ , opens the signal gate and the counting chain starts operating. On the second pulse from the time-base, spaced one second after the first, the output from  $V_1$  again feeds via  $C_3$  to binary stage  $V_3$ , which changes back to its original state as well as sending a pulse to operate the latching binary  $V_4$ . This immediately changes over and in so doing the right-hand anode goes negative driving the suppressor grid of  $V_1$  negative via the left-hand cathode-follower  $V_2$ , and so stops any further time-base pulses getting through time-base gate  $V_1$ . In changing back to its original state  $V_3$  drives the suppressor grid of the signal gate valve ( $V_3$  in Fig. 5) negative via the right-hand cathode follower  $V_2$  and closes the gate. Thus the signal has been allowed

Fig. 7. Crystal oscillator and first Phantastron divider



to get through the signal gate for one second only. The figures now displayed on the counting chain indicate the pulses per second, or in other words, frequency in digital presentation. Plug-in construction is also used, see Fig. 6.

The displayed frequency may be indicated until the reset control is manually operated, or alternatively an automatic recycling circuit may be used. The right-hand anode of

ing and stopping the counter chain (opening and closing signal gate), a clean pulse is presented to the appropriate binary stage in the control circuits.

#### POWER UNIT

This is quite conventional using a hard vacuum rectifier valve for all h.t. and bias supplies.

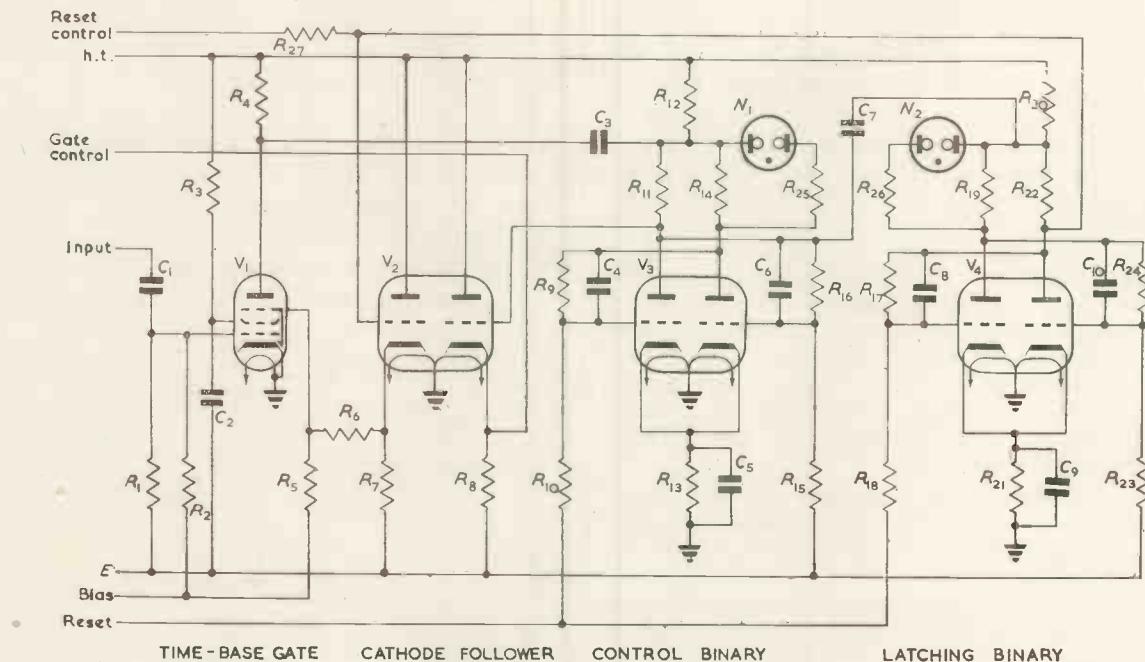


Fig. 8. Control circuits

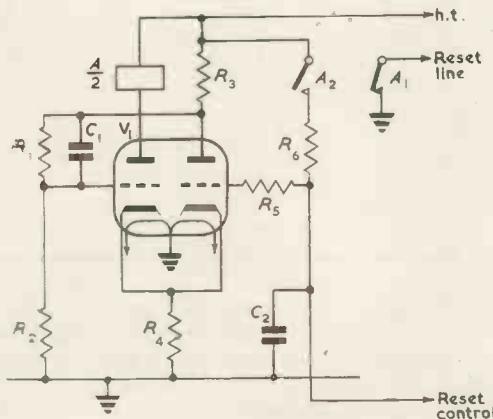


Fig. 9. Automatic reset circuit

V<sub>4</sub> goes negative on the receipt of the second pulse from the time-base and this potential is fed via a high resistance to the right-hand grid of a double triode, Schmidt trigger circuit V<sub>1</sub>, in Fig. 9. C<sub>5</sub> is initially charged sufficiently positive to make the right-hand triode of V<sub>2</sub> conducting and the left-hand triode non-conducting, consequently relay A is not energized. The capacitor C<sub>5</sub> is slowly discharged when the latching binary (V<sub>4</sub> in Fig. 8) changes over. The right-hand anode of V<sub>1</sub> in Fig. 9 slowly rises negatively until such time as the triggering action takes place, energizing relay A, which opens the reset line via contacts A/1 to all the appropriate units, and charges C<sub>2</sub> via A/2 and R<sub>6</sub> ready for the next operation. An additional trigger valve in the control circuit is used in conjunction with the push buttons, so that when they are operated for manually start-

#### Conclusions

Apart from measuring frequencies in the range 10c/s to 100kc/s the instrument may be used for many other purposes. For instance, with the aid of an auxiliary tachometer generator, speed measurements of revolving shafts may be made. If the generator is designed to provide 60 pulses for each revolution, a direct reading of revolutions per minute will be obtained during the one-second counting period of the instrument. Another useful feature for the accurate measurement of low frequencies and time intervals may be obtained by interchanging the signal and time-base channels. Thus, by permitting the 100kc/s oscillations from the crystal to enter the signal channel and feeding the unknown pulses into the time-base channel, a direct time measurement between successive pulses, in discrete 10μsec increments, will be indicated. In fact, for low frequency measurement, below about 300c/s, this is the most accurate method to employ. As an example, in measuring the mains frequency at 50c/s, an accuracy of 0.05 per cent is achieved.

The portability of the instrument makes it a useful tool on the production line, particularly for checking calibration of oscillators, filters and similar frequency sensitive apparatus.

#### REFERENCES

1. ECCLES, W. H., JORDAN, F. W. *Radio Rev. I*, 143 (1919).
2. FERGUSON, G. J., FRASER, G. H. The Design of Four Decade Tube Scalers. *Rev. Sci. Instrum.* 22, 937 (1951).
3. GROSDOFF, I. E. Electronic Counters. *RCA Rev.* 7, 428 (1946).
4. GOTTLIEB, I. Decade Counter. *Wireless World* 60, 234 (1954).
5. POTTER, J. T. A Four-Tube Counter Decade. *Electronics* 17, 110 (June, 1944).
6. Waveforms, p. 611 (McGraw-Hill).
7. BUTLER, F. Cathode Coupled Oscillator. *Wireless Engr.* 21, 521 (1944).
8. Waveforms, p. 577 (McGraw-Hill).

# Chrominance Circuits for Colour Television Receivers

(Part 1)

By B. W. Osborne\*, M.Sc., A.Inst.P., A.M.Brit.I.R.E

*A survey of current practice and recent developments in phase synchronization, chrominance demodulator and matrix circuits for use in colour television receivers.*

THE specifications for colour television signals<sup>1</sup> formulated by the National Television System Committee (N.T.S.C.) and approved by the Federal Communications Commission of the U.S.A. have formed a sound basis for the establishment in that country of a commercially practicable colour television service, completely compatible with existing black and white transmissions, and which does not require additional frequency allocations. The N.T.S.C. "constant luminance" system is now well known.

Though at the time of writing there are no British standards for colour television transmissions, a proposal has been published for the adaptation of the N.T.S.C. standards to the 405-line system<sup>2</sup>.

This article has been written in the form of a brief review of circuit techniques applicable to the chrominance section of a colour television receiver designed to receive an N.T.S.C. type signal.

In the N.T.S.C. system<sup>1</sup>, the chrominance information is carried on a suppressed sub-carrier operating in the upper part of the video frequency band. For optimum performance within a limited bandwidth, this sub-carrier is modulated by two colour difference signals in phase quadrature. These colour difference components, which are termed the  $E_I'$  and  $E_Q'$  signals, are combined at the transmitter to form the chrominance signal. Thus the instantaneous phase of the chrominance signal varies according to the relative amplitudes of the  $E_I'$  and  $E_Q'$  signals (Fig. 1), and therefore, as  $E_I'$  and  $E_Q'$  are colour difference signals, according to the instantaneous chromaticity.

The complete colour signal  $E_M$  may then be described, as the sum of the luminance and chrominance signals, by the following expression<sup>1</sup>:

$$E_M = E_Y' + [E_Q' \sin(\omega t + 33^\circ) + E_I' \cos(\omega t + 33^\circ)],$$

where  $E_I' = -0.27(E_B' - E_Y') + 0.74(E_R' - E_Y')$ ,

$$E_Q' = 0.41(E_B' - E_Y') + 0.48(E_R' - E_Y'),$$

and the luminance signal by:

$$E_Y' = 0.30E_R' + 0.59E_G' + 0.11E_B'.$$

In these equations  $E_R'$ ,  $E_G'$  and  $E_B'$  are the voltages, corrected for display tube non-linearity<sup>3</sup>, to be applied to the red, green and blue guns respectively, of a colour display tube with suitable phosphor characteristics (C.I.E. chromaticities (0.67, 0.33) for red, (0.21, 0.71) for green and (0.14, 0.08) for blue).

It follows that  $E_R'$ ,  $E_G'$ , and  $E_B'$  may be obtained in the chrominance section of a receiver by suitably matrixing the  $E_I'$ ,  $E_Q'$  and  $E_Y'$  channels.

Both the choice of axes for the  $E_I'$  and  $E_Q'$  signals with respect to the horizontal ( $E_B' - E_Y'$ ) axis and the vertical ( $E_R' - E_Y'$ ) axis (Fig. 1), and the choice of relative frequency bandwidths for the  $E_I'$  and  $E_Q'$  channels, are governed for optimum picture quality by the known characteristics of the human eye, as well as by the practical limita-

tions inherent in the problem of transmitting  $E_Y'$ ,  $E_I'$  and  $E_Q'$ , independently within a limited frequency band<sup>4,5,6</sup>.

These factors lead to the use of:

(a) A full video bandwidth signal (the  $Y$  channel) giving shades of grey for fine picture detail.

(b) A moderately wide band colour mixture signal (the

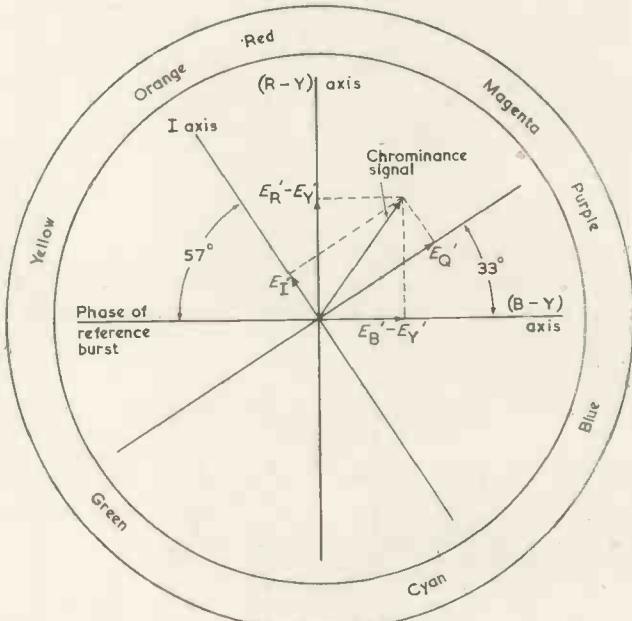


Fig. 1. Phase relationships in the N.T.S.C. chrominance signal

$Y$  channel) distinguishing orange-red from blue-green, and thus providing two colour primaries for small objects.

(c) An additional narrow-band colour mixture signal (the  $Q$  channel) which in conjunction with the  $I$  and  $Y$  signals effectively provides three colour primaries for large objects.

The frequency and phase response required in the receiver circuits for the  $Y$ ,  $I$  and  $Q$  channels will be discussed later.

In order to determine the instantaneous chromaticity, it is necessary to determine the instantaneous phase difference between the chrominance signal and the phase reference. The latter, at sub-carrier frequency, is transmitted as a short burst occurring during the line flyback period.

To obtain the colour information, it is then necessary to obtain a local c.w. phase reference, synchronized to the transmitted phase reference bursts. Various methods of achieving the high degree of frequency and phase stability required will be described.

The nature of the circuits following the second detector in a colour television receiver is indicated by the simplified block diagram of Fig. 2. Referring to this figure, it is seen that no less than four separate outputs are taken from the

\* Ultra Electric Ltd.

video stage. One feeds a normal sync separator circuit, which controls the initiation time of the line and frame scans in the usual manner. The second provides the burst output for phase synchronization and is fed to a narrow band sub-carrier frequency amplifier, gated to operate only during the line flyback period when the burst is present. The burst output controls an automatic phase control system, which provides an output at sub-carrier frequency locked in the desired phase relationship to the transmitted burst.

The third and fourth outputs taken from the video circuit provide the chrominance and luminance information respectively. The former, carried on the sidebands of the sub-carrier in the upper region of the luminance spectrum, is fed through a suitable band-pass filter and amplifier to two synchronous demodulators. These demodulators receive c.w. signals in phase quadrature, aligned in phase for the *I* and *Q* (or the *R-Y*, *B-Y*) axes (Fig. 1), from the phase reference source. By the use of suitable

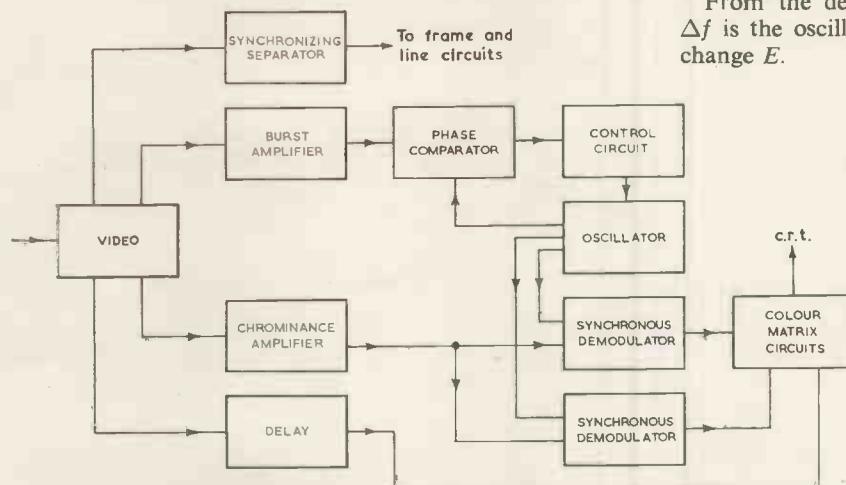


Fig. 2. Block diagram of the luminance and chrominance circuits of a typical television receiver

low-pass filters, the *I* and *Q* (or the *R-Y*, *B-Y*) signals are obtained separately from the two demodulator output circuits, and are then added in the correct sense and proportion to the delayed luminance signal, in order to obtain the  $E_R'$ ,  $E_G'$  and  $E_B'$  signals to be applied to the red, green and blue guns of the display tube.

A comparison of the (*I*, *Q*) and the (*R-Y*, *B-Y*) demodulator systems is included in the discussion of synchronous demodulator and matrix circuits.

#### Phase Synchronization of the Local Oscillator

##### THE AUTOMATIC PHASE CONTROL LOOP

The basic system for phase synchronization is illustrated in Fig. 3. The phase detector compares the phase of the local oscillator with that of the incoming signal, and generates an output voltage which varies in amplitude according to the phase difference and in sense according to whether the frequency of the local oscillator lies above or below that of the incoming signal.

This d.c. phase detector output is fed through a low-pass filter to a reactance valve and oscillator, in the correct sense for the reduction of any initial frequency or phase difference between the local oscillator and the incoming burst.

The feedback loop has a pull-in performance that is dependent on the loop gain and on the loop frequency response. Following Richman<sup>7</sup>, we may describe the performance of an automatic phase control (a.p.c.) loop using the following definitions:

(1) Let the rate of change of the oscillator frequency with the applied reactance valve output voltage be  $\beta$ .

(2) Let the rate of change of the phase detector output voltage  $E$  with the phase deviation  $\phi$  be  $\mu$  volts per radian at the operating point ( $\phi = 0$ ).

(3) Let the gain of the low-pass filter at high frequencies be  $m$ , where  $m \ll 1$ , compared with a value of unity for d.c.

The d.c. loop gain is then  $\mu\beta$ . At frequencies high compared with the noise bandwidth of the low-pass filter, the a.c. loop gain is  $m\mu\beta$ , a much lower value than  $\mu\beta$ .

The static phase error of the loop,  $\Delta\phi$ , is given approximately by the phase detector characteristic in conjunction with the definition of  $\mu$  and  $\beta$  given above. In the phase detector circuit, a small change  $\Delta\phi$  in the phase-angle between the two applied signals produces a small change  $\Delta E$  in the voltage applied to the reactance valve, where  $\Delta E = \mu \sin \Delta\phi$ .

From the definition of  $\beta$  we have  $\Delta E = \Delta f / \beta$ , where  $\Delta f$  is the oscillator frequency shift produced by a voltage change  $E$ .

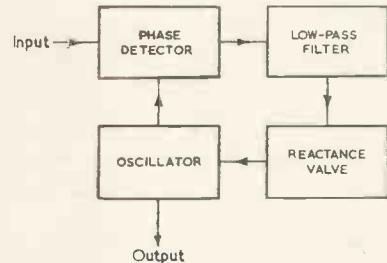


Fig. 3. The basic automatic phase control loop

Thus:

$$\sin \Delta\phi = \Delta E / \mu = \Delta f / \mu\beta \dots \dots \dots (1)$$

High values of loop gain thus give tight static phase control.

To take a numerical example, if the possible overall oscillator frequency drift, due to all causes, is 3kc/s, then for the limit  $\Delta\phi < 5^\circ$  (set by considerations of colour reproduction in the N.T.S.C. system as greater phase change may produce noticeable colour changes), we require a value of  $\mu\beta$  greater than 34 000.

Values of  $\beta$  as great as 100 000 have been obtained at 2.6Mc/s using a normal reactance valve circuit employing an EF92 valve, though such high values of  $\beta$  are not generally compatible with high long-term stability. The effective value of  $\beta$  can be increased if necessary by the use of a d.c.-coupled degenerative amplifier stage between the phase detector and the reactance valve, but precautions must be taken to ensure that the gain of such a stage remains constant and independent of supply voltage changes.

The determination of  $\mu$  for certain phase detector circuits is discussed later. Practical values of  $\mu$  generally lie within the range from 5 to 150V/rad.

The time taken for an automatic phase control system to pull in and to lock in phase is dependent on the noise bandwidth of the system<sup>7</sup>, and this depends on the characteristics of the filter following the phase detector. This

filter will generally take the form of the circuit of Fig. 4, where the ratio  $m$  of a.c. to d.c. is equal to  $x/(1+x)$ , and is independent of the values assigned to  $R$  and  $C$ . For  $x \ll 1$ , we have  $m \approx x$ .

Richman<sup>7</sup> has shown that the equivalent noise bandwidth of an a.p.c. loop is then given approximately, in terms of the filter characteristics, by:

$$f_n = (\pi/2) [m\mu\beta + (1/2\pi xRC)] \dots \dots \dots (2)$$

In an alternative form, substituting in terms of a damping factor  $K$ , where  $K = (\pi/2) m\mu\beta xRC$ , we have:

$$f_n = (\pi/2) m\mu\beta \cdot \frac{(K + \frac{1}{4})^3}{K} \dots \dots \dots (3)$$

The optimum conditions for dynamic "pull-in" performance have been investigated by Richman<sup>7</sup>, who has shown that the optimum value for  $K$  is  $K = \frac{1}{2}$ . For operation well within the frequency range over which the system will lock, the minimum frequency pull-in time  $T_f$  is then given in terms of the noise bandwidth by the relationship:

$$T_f = [(\pi/2K) \cdot \Delta f]^2 \cdot \frac{(K + \frac{1}{4})^3}{f_n} \text{ or } T_f > 4[(\Delta f)^2/f_n^3]$$

i.e.:

$$\Delta f < \frac{1}{2}(T_f)^{1/2} (f_n)^{3/2} \dots \dots \dots (4)$$

where  $\Delta f$  is the initial frequency difference.

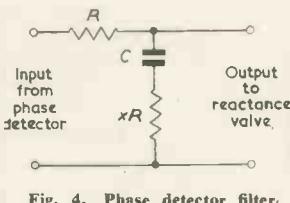


Fig. 4. Phase detector filter.

An additional limitation in the case of a "burst" synchronizing signal gated at the line frequency  $f_h$  is that  $\Delta f \leq f_h/2$ .

The optimum value for the noise bandwidth  $f_n$  lies in the region of  $100c/s^7$ . Taking this value of  $f_n$ , we see that for a maximum tolerable pull-in time of one second, the oscillator frequency deviation must not exceed about  $500c/s$ .

Thus the use of a simple a.p.c. loop, in conjunction with an LC oscillator, would necessitate the provision of a manual control of oscillator frequency in order to obtain a phase lock. The frequency range over which such a system will lock can be extended (up to  $f_h/2$  for a gated signal) by increasing  $\mu\beta$  and decreasing  $m$ , but the pull-in time becomes excessive. This disadvantage has been overcome in a modified form of a.p.c. system due to Richman<sup>8</sup>, which employs two synchronous detectors operating in quadrature.

Alternatively, if the maximum oscillator frequency deviation is limited to a few hundred cycles per second, it is obviously possible to obtain the desired phase stability using a simple a.p.c. loop. An example of the design of an a.p.c. loop to be used in conjunction with a crystal-controlled oscillator is given later.

#### PHASE DETECTORS

The function of a phase detector in an a.p.c. loop of a colour television receiver is to compare the instantaneous phase of the local reference oscillator with that of the incoming sub-carrier burst during the line flyback period, and to provide a d.c. output voltage dependent in sense and amplitude on the phase difference. This output voltage is then used to control the frequency and phase of the

local reference oscillator so that it remains locked in phase to the sub-carrier burst.

For optimum performance, and to eliminate steady state phase errors, the following conditions are desirable:

(a) The balance point should be independent of burst amplitude.

(b) The balance point should be independent of any fluctuations in the applied voltages, or in the power supply lines.

(c) The phase detector characteristic should be approximately linear and symmetrical about the balance point.

(d) The gradient of the phase detector characteristic at the balance point should be as high as possible.

The sensitivity of the phase detector is measured in terms of the gradient  $\mu$  of the phase detector characteristic at

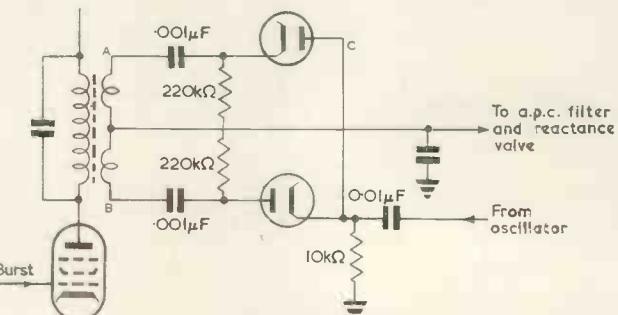


Fig. 5. A common-phase detector circuit

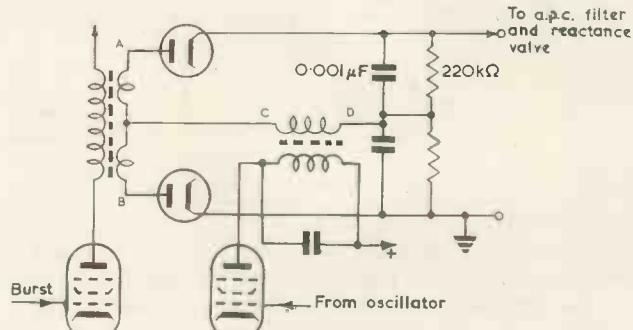


Fig. 6. An alternative phase detector circuit

the balance point, which must be known in order to determine the characteristics of the phase control system as a whole.

A commonly used phase detector circuit is that of Fig. 5. In this circuit the incoming burst appears across the centre-tapped secondary winding AB, so that the burst signal applied to the anode of one diode is in anti-phase with that applied to the cathode of the other. The signal from the local reference oscillator is applied at C, where its amplitude should be about equal to half the voltage across AB. When the phase of the local oscillator signal at C is in quadrature with the burst, the system is theoretically balanced and the resultant rectified output applied to the reactance valve is zero. Any change in the phase of the oscillator signal at C will produce a d.c. output to the reactance valve, the polarity of which reverses as the phase at C passes through the balance point.

An alternative circuit is that of Fig. 6, which has been discussed by Farren<sup>9</sup>. As in the previous circuit, the burst signal voltage across AB should be double the oscillator voltage across CD. At the balance points the signals are in quadrature.

The performance of the circuit of Fig. 6 was compared with that of Fig. 5, using similar GE35 germanium detectors and the same applied voltages in both circuits. The value of  $\mu$  at the balance point for the circuit of Fig. 6 was found to be approximately double that of Fig. 5. In order to simulate conditions in a colour television receiver a.p.c. loop, the measurements were repeated with both the signals gated at the line frequency, the signals being applied to the phase detector for  $5\mu\text{sec}$  during each period of  $100\mu\text{sec}$ . For both circuits this resulted in a reduction of the value of  $\mu$  by a factor of about 0.7.

In practice, values of  $\mu$  of up to about  $40\text{V}/\text{rad}$  may be obtained with circuits similar to that of Fig. 6, but at the high levels it is not easy to maintain the symmetry of the phase detector characteristic.

With either of the nominally balanced circuits of Figs. 5 and 6, any unbalance will produce an asymmetrical pull-in characteristic. The exactness of phase balance affects both the long-term stability of the system and the

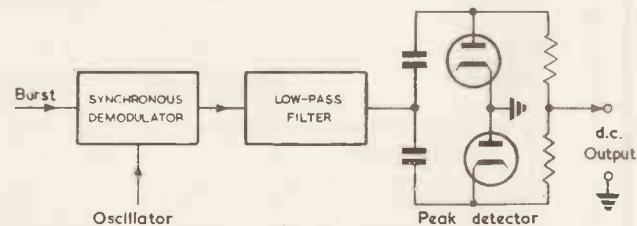


Fig. 7. Bi-polar phase detector

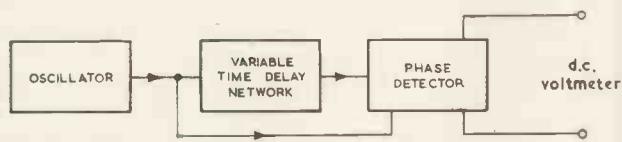


Fig. 8. Measurement of phase detector characteristic

sensitivity. Poor long-term stability may necessitate the frequent use of a master phase control in a colour television receiver in order to obtain satisfactory colour reproduction.

The upper limit of phase detector sensitivity is set, for a given circuit, by the available input voltage, and this must be limited to avoid unwanted feedback. The most stable phase detector system is that which only employs low impedance low voltage circuits, and which consequently will only give low values of  $\mu$ .

A self-balancing phase detector of high stability and sensitivity has recently been described by Clark<sup>10</sup>. A gated synchronous detector is used in conjunction with a bi-polar detector circuit. The load of the synchronous detector consists of a low-pass filter accepting the burst envelope, the sense and amplitude of which depends on the direction and magnitude of any phase deviation. The bi-polar detector consists of two peak detectors whose outputs are combined, as in Fig. 7. The d.c. output is then equal to half the difference between the positive and negative peaks of the pulse train applied. Zero d.c. output at the balance point is assured without the need for using any preset controls.

As the gain of the system is at burst frequency, the sensitivity may be increased by inserting a linear voltage amplifier between the synchronous and bi-polar detectors, without any risk of instability. The use of a synchronous demodulator also permits considerable variation in the amplitude of the input signals without impairing the circuit performance.

Alternative arrangements<sup>10</sup> include the use of a double-triode cascode demodulator circuit, with the input burst and the local reference oscillator signals applied separately to the grids. A frequency selective circuit between the two series triodes is then used to remove the line frequency components imposed in the first triode before demodulation in the second.

Another phase detector described by Clark<sup>10</sup> using a double-triode clipping amplifier between the demodulator and the bi-polar detector gave a sensitivity as high as  $150\text{V}/\text{rad}$ .

The characteristic of a phase detector may be measured by an oscillator and a calibrated variable time-delay network, as in Fig. 8. The variable delay most conveniently takes the form of a tapped delay line, which may be calibrated directly in phase at the sub-carrier frequency. A suitable line designed and used for this purpose consists of 36 sections of characteristic impedance  $540\Omega$ , and has constant delay characteristics over the video spectrum (i.e. a high cut-off frequency of about  $80\text{Mc/s}$ ). This line is fed from a cathode-follower matched to the line at the input end, and is terminated by a carbon resistor. Each step corresponds to approximately  $15^\circ$  phase delay at  $2.66\text{Mc/s}$ . The output is obtained through a cathode-follower mounted on a bridge which may be traversed along the line. The system thus constitutes a stable 36-position switched time-delay, with the inherent advantage of being usable at any frequency within the video band.

Precision calibrated single-frequency phase-shifters of high performance have been described by Morrison, Karstad and Behrend<sup>11</sup>, and by Stern<sup>12</sup>.

#### THE QUADRICORRELATOR Two-MODE CONTROL CIRCUIT

In a simple a.p.c. loop, the requirements for good dynamic phase stability conflict with those for rapid pull-in. Richman<sup>7,8</sup> has developed a two-mode automatic phase control system which has good characteristics for rapid pull-in when the reference oscillator is not in synchronism with the burst, and good dynamic stability, with consequently poor pull-in performance, when in synchronism.

This performance is achieved by the addition of a second phase detector to the a.p.c. loop, which, following Richman, we will call the *S* detector. The *S* detector operates in quadrature with the  $\phi$  detector, and thus produces a unidirectional d.c. output when the system is in synchronism. This output is used to control the a.c. gain of the  $\phi$  circuit, the a.c. gain being least when the oscillators are in synchronism, giving low noise bandwidth and hence high phase stability, and greatest when the oscillators are not in synchronism, for rapid pull-in (see equations (2) and (3)).

The balance point of the  $\phi$  detector circuit corresponds to  $90^\circ$  phase difference between the burst signal and the reference oscillator signals. The reference oscillator voltage fed to the *S* detector must, therefore, be shifted in phase by  $90^\circ$  to obtain the desired operating conditions.

A further advantage of this two-mode system is that the *S* detector output is available for the control of the chrominance amplifiers. The chrominance amplifiers can then be biased off when no output is obtained from the *S* detector, i.e., when no burst signal is received. The chrominance circuits may thus be rendered inoperative during the reception of monochrome transmissions. This is often referred to as "colour killer" action.

A detailed discussion of various forms of quadricorrelator is given by Richman<sup>8</sup>, who has shown that frequency pull-in ranges of  $\pm 5\text{kc/s}$  may be easily obtained with a pull-in time of less than one second. Furthermore, the system maintains its performance under conditions of very low signal-to-noise ratio.

A typical quadricorrelator system is illustrated in Fig. 9. The basic a.p.c. loop consists of the  $\phi$  detector, the a.p.c. filter, the reactance valve and the oscillator. A gated amplifier receives the burst-signal from a 2.66Mc/s trap in the video amplifier, and provides an output of some 10V or 20V r.m.s. The gating waveform is generally derived from a secondary winding on the line output transformer and allows the valve to conduct during the line flyback period when the burst is received. The burst output is fed to the control grids of the two synchronous demodulators, the quadrature outputs from the reference oscillator being fed to the two suppressor grids. The  $\phi$  detector output is a.c. coupled through the quadricorrelator amplifier to the reactance valve, by-passing the low-noise-bandwidth a.p.c. filter. The  $S$  detector output is fed through a low-pass filter and controls the bias applied to the quadricorrelator amplifier, which is allowed to conduct when the signals are out of synchronism. The  $S$  detector output is also used to control

determines the phase of the synchronized oscillator relative to the burst. In practice, the master phase control is set to give correct colour rendition.

The two-mode quadricorrelator may, with advantage, be used with self-balancing  $\phi$  and  $S$  phase detectors consisting of synchronous demodulators and bi-polar detectors, as described previously. A linear amplifier stage consisting of a triode with current feedback can then be connected between the  $\phi$  demodulator and its associated bi-polar detector in order to increase the sensitivity of the  $\phi$  circuit<sup>10</sup>.

The  $\phi$  detector output is fed to the reactance valve through a filter of the type shown in Fig. 4.

The output from the  $S$  detector is fed through a low-pass filter (to improve system characteristics in the presence of noise) to the d.c.-coupled beat note amplifier. The output of this stage is a.c.-coupled to the reactance valve and d.c.-coupled to the colour killer circuit, and may also be used to control the gain of the a.p.c. loop.

#### THE USE OF A CRYSTAL-CONTROLLED OSCILLATOR IN AN A.P.C. LOOP

To obtain a pull-in time of less than one second with static phase errors not exceeding  $\pm 5^\circ$  in an a.p.c. loop designed for optimum dynamic performance, the local oscillator frequency deviation must never exceed about 0.02 per cent of the sub-carrier frequency, i.e.,  $\pm 500\text{c/s}$  at 2.66Mc/s. In the case of a simple *LC* oscillator circuit, this high stability may not be realized, so that a manually operated oscillator frequency control would then have to be provided.

We have seen, however, that the addition to the a.p.c. loop of a second phase detector system, working in quadrature with and controlling the characteristics of the main feedback

loop, makes it possible to design a satisfactory phase synchronization system using a simple *LC* oscillator, at the expense of an increase in the number of valves in the a.p.c. circuit.

An alternative and simpler system, which has found favour in several commercial designs, is that in which the basic a.p.c. loop is used in conjunction with a crystal-controlled oscillator, the maximum initial frequency deviation of which can be limited to about 0.005 per cent. It is then possible to choose the loop parameters so as to obtain tight static phase control.

From equation (1), taking  $\Delta f = 150\text{c/s}$  and  $\Delta\phi = 5^\circ$ , we require a value for  $\mu\beta$  of about 1700. When controlling the frequency of a crystal oscillator circuit with an ordinary pentode reactance valve such as the EF92, values of  $\beta$  of about 50cycles per volt may be obtained. However with a high slope valve such as the new Mullard E180F pentode, a higher value of  $\beta$  should be obtainable.

Taking  $\beta = 50$ , for  $\mu\beta = 1700$  we have  $\mu = 34\text{V/rad.}$ , a phase detector sensitivity which is easily realized with the circuit of Fig. 6.

Referring to equation (3), taking  $K = \frac{1}{2}$  and  $f_n = 100\text{c/s}$ , we see that for optimum performance  $m\mu\beta = 400/3\pi$ . For  $\mu\beta = 1700$ , we then have  $m = 0.024$ . It follows that the filter circuit associated with the phase detector output must be designed for this ratio of a.c. to d.c. gain.

The burst signal, obtained from a 2.66Mc/s trap in the anode circuit of the first video amplifier, is fed through gated amplifier and limiter stages to the phase detector, where the burst phase is compared with that of the signal

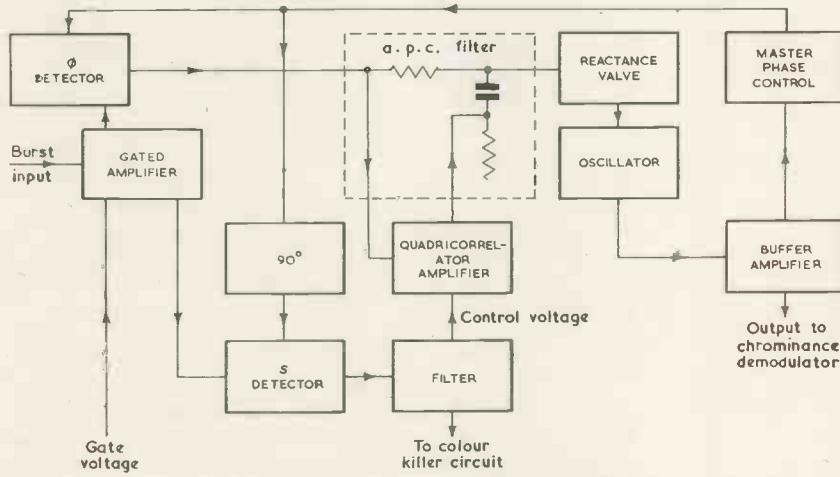


Fig. 9. A quadricorrelator phase control system

the gain of the a.p.c. loop amplifier stages, the gain of the loops being reduced in the synchronous condition. Richman<sup>8</sup> coupled the screen grids of the valves in the a.p.c. loop to those in the chrominance amplifier, using a common voltage dropping resistor. The chrominance amplifier is cut off by the colour killer control voltage when the system is out of synchronism, so that a higher voltage is then applied to the screen grids of the valves in the a.p.c. loop. The gain of the latter is thus increased when the reference oscillator is not synchronized to the burst. Short suppressor base valves such as the 6AS6<sup>13</sup> or 6F33 types are suitable for use as demodulators.

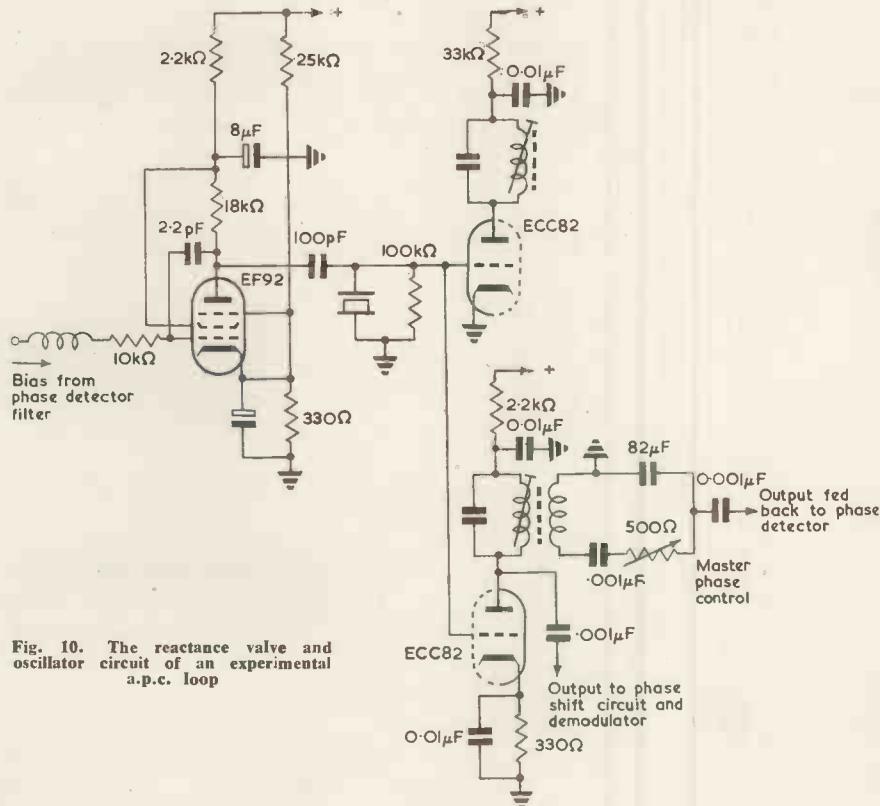
The reactance valve and oscillator circuit follows normal practice. The value of  $\beta$ , the rate of change of frequency with applied voltage, depends on the fixed capacitance across the tuned circuit and on the static bias applied to the reactance valve. Using an EF92 valve with 4.5V bias, and with a fixed 20pF capacitor across the tuned circuit, the value of  $\beta$  as measured under static conditions was found to be approximately 20kc/s/V. For stability, higher values of  $\beta$  are undesirable. A by-passed self-bias resistor in the reactance valve cathode circuit may be used to provide d.c. degeneration, thus reducing the effect of long-term changes in the reactance valve characteristics.

The oscillator output is fed through a buffer stage to a master phase control and thence to a quadrature transformer which provides the sine and cosine outputs to the  $\phi$  and  $S$  phase detector circuits. The oscillator output to the chrominance demodulators is taken directly from the buffer stage, so that the setting of the master phase control

fed back from the crystal oscillator. The phase detector output voltage controls the oscillator phase by means of the reactance valve, as in the circuit of Fig. 10. The phase of the oscillator output relative to the burst may be preset by use of the master phase control shown in the figure, in order to provide an output of the phase required for correct operation of the colour demodulators.

## CRYSTAL RINGING CIRCUITS

Methods of synchronizing a local oscillator to the burst signal, in order to generate a continuous wave phase reference, have been discussed above. An alternative means of obtaining the local phase reference signal is to use the burst to excite a high Q crystal circuit, so that oscillations in the correct phase are maintained



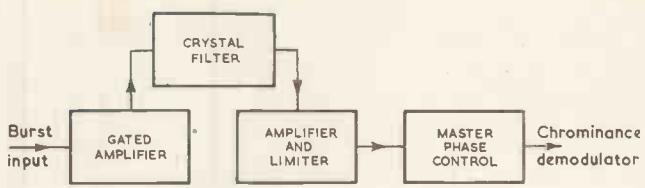
**Fig. 10.** The reactance valve and oscillator circuit of an experimental a.p.c. loop

during the line period until the arrival of the next burst.

This crystal ringing system provides an attractive alternative to the use of an a.p.c. loop, as it uses fewer valves and components, and is capable of comparable performance. However, the circuit adjustments are critical and a higher degree of crystal frequency stability is required than when, say, a crystal-controlled oscillator is used in an a.p.c. loop.

The system may be regarded as a narrow band crystal filter which only passes the fundamental frequency component of the synchronizing burst. If the filter, considered as a single parallel resonant circuit, is driven from a constant-current source (during the burst interval), the static phase error of the system is given in terms of the frequency deviation  $\Delta f$  by  $\Delta\phi = \tan^{-1} 2Q \Delta f/f$ .

The higher the Q of the ringing circuit, the less will it be influenced by the presence of noise. Furthermore, a very high Q ringing circuit has an inherently long voltage decay time-constant. However, an upper limit to the desirable circuit Q is set by the possibility both of changes in the crystal resonant frequency and of small changes in



**Fig. 11.** A crystal ringing phase reference generator

the sub-carrier frequency at the transmitter. Values of Q of between 5 000 and 10 000 are generally acceptable. As the Q of the crystal itself is appreciably higher, there is no practical difficulty in obtaining the desired Q value, but the tuning of the crystal filter is extremely critical.

A crystal ringing synchronization system (Fig. 11) may consist simply of a gated burst amplifier with the crystal in a series resonant mode coupled to the anode circuit, followed by two amplifier stages, the second of which acts as a limiter and feeds signals of constant amplitude to the colour demodulators.

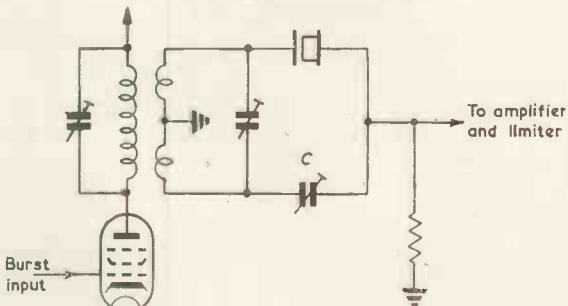
Synchronization systems of this type have been discussed by Good<sup>14</sup>, who used an AT-cut quartz crystal in a bridge type circuit in order to neutralize the shunt capacitance (as in Fig. 12, where the capacitor  $C$  is the neutralizing component). Neutralization is obtained by adjusting  $C$  so as to reduce the transient occurring during the presence of the burst. Good, who used a  $Q$  of 8 000 for the N.T.S.C. signal, found that it was advisable to contour grind the crystals in order to avoid the production of striations, due to the existence of spurious modes separated in frequency from the fundamental by either 3 or 5 times the line frequency<sup>14</sup>.

# THE USE OF ALLOY JUNCTION TRANSISTORS

An experimental circuit using a single transistor and crystal which may be capable of satisfactorily replacing the phase detector, reactance valve and crystal oscillator, i.e., the entire automatic phase control loop, of a conventional colour receiver, has been described by Kabell and Evans<sup>15</sup>.

barrier capacitance is inversely proportional to the square-root of the collector-to-base voltage. Thus the output capacitance may be controlled by varying the collector voltage. This characteristic permits the use of the transistor in an automatic phase control circuit.

**Fig. 12.** A crystal filter circuit (after Good)



In the self-running oscillator circuit shown in Fig. 13, the feedback current passing from emitter to base during the burst interval will be altered by an amount depending on the phase difference between the incoming burst signal and the oscillator feedback signal. If the phase difference is such that the emitter current is increased during the burst,  $C_2$  will be slightly discharged and the collector voltage will be decreased. The output capacitance will therefore increase, lowering the resonant frequency of the tuned circuit and changing the reactive component of the generator impedance feeding the series-resonant crystal filter. The crystal frequency is thus changed so as to synchronize the oscillator with the burst.

The circuit locks with the phase of the feedback current at 90° to the burst.

Kabell and Evans obtained a rate of change  $\mu'_1$  of output capacitance  $\Delta C$ , with phase difference  $\phi$ , for a suitable alloy transistor, of approximately 1.14 pF/rad.

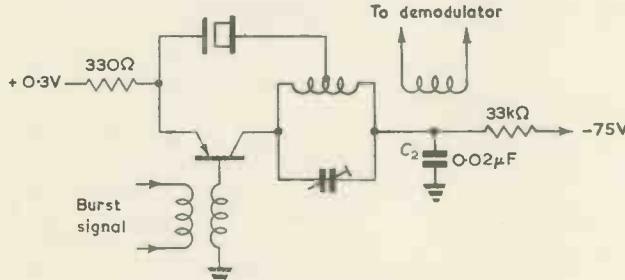


Fig. 13. Transistor a.p.c. loop

The rate of change of control reactance with transistor output capacitance change,  $\beta_1$ , is given by:

$$\beta_1 = \frac{\partial X}{\partial (\Delta C)} = -j\omega n(\omega LQ)^2,$$

where  $X$  is the reactance appearing in series with the crystal and  $n$  is the impedance step-down ratio of the tap on the inductance  $L$ .

For  $n = 0.05$  and  $\omega LQ = 21\text{k}\Omega$  for 3mW output, a value of  $\beta_1$  of  $494\Omega/\text{pF}$  can be realized at  $3.58\text{Mc/s}$ .

The rate of change of crystal resonant frequency with external series reactance at the series resonant frequency of the crystal is given by:

$$\beta_2 = (\partial f / \partial X) = (\omega^2 C_1 / 4\pi) = 40 \cdot 10^{12} C_1,$$

## A New Automatic Announcer

The General Electric Co. Ltd has developed a novel automatic announcing system which can reproduce, at the touch of a button, any one of 25 pre-recorded announcements or signals, which can then be relayed over a loudspeaker system.

The messages are recorded side by side on a special loop of 3in wide magnetic tape, and the appropriate one selected by push-button when required. To make the system fully automatic, particular announcements such as shift changes, meal breaks or other messages can be triggered off by a master clock mechanism without the need for any human attendance. If required, the push-buttons can be mounted remotely from the instrument.

The automatic announcer was originally developed by the G.E.C. for British Railways (Eastern Region) and first installed at Stratford Station. In this installation the platform announcements for passengers are automatically initiated by the trains themselves, first as they approach the station and then again as they come to a standstill at the platform. Equipment of this type has now been adapted for general industrial use and may be applied in any situation where numbers of routine broadcasts have to be repeated fairly frequently.

where  $C_1$  is the equivalent internal series capacitance of the crystal. It is therefore desirable to use a crystal with as large a value of  $C_1$  as possible.

The loop gain is then given by:

$$(\partial f / \partial \phi) = \mu' \beta_1 \beta_2.$$

Kabell and Evans obtained a value for  $\beta_2$  of 1.66 cycles/ohm, and for  $\partial f / \partial \phi$  and for  $\partial f / \partial \phi$  of 935 cycles/radian.

In other words, for a maximum tolerable static phase error of  $\pm 5^\circ$ , the necessary tolerance on the free-running oscillator frequency is  $\pm 82\text{c/s}$ , which should be obtainable with reasonable precautions. A negative temperature coefficient capacitor should be used as part of the tuned circuit. If the transistor base is returned to a tap on a potentiometer chain across the collector supply, the performance of the circuit is less dependent on supply voltage variations.

A suitable transistor for use in colour receiver a.p.c. loops must be capable of about 15dB gain at 2.66Mc/s and should have a relatively large collector capacitance ( $\sim 20\text{pF}$ ) with a low value of  $I_{co}$ . (The Raytheon 2N112 has been found suitable.)

If the simple transistor a.p.c. loop can be developed to a stage where it can be manufactured commercially, it will considerably simplify the decoder circuit of colour receivers.

## REFERENCES

1. N.T.S.C. Signal Specifications. *Proc. Inst. Radio Engrs.* 42, 17 (1954).
2. Marconi Company 'Experimental Colour Television.' *J. Televis. Soc.* 7, 236 (1954).
3. LIVINGSTON, D. C. Colorimetric Analysis of the N.T.S.C. Colour Television System. *Proc. Inst. Radio Engrs.* 42, 138 (1954).
4. BROWN, G. H. The Choice of Axes and Bandwidths for the Chrominance Signals in N.T.S.C. Colour Television. *Proc. Inst. Radio Engrs.* 42, 58 (1954).
5. BAILEY, W. F., HIRSCH, C. J. Quadrature Cross-talk in N.T.S.C. Colour Television. *Proc. Inst. Radio Engrs.* 42, 84 (1954).
6. ABRAHAMS, I. C. The Frequency-Interleaving Principle in the N.T.S.C. Standards. *Proc. Inst. Radio Engrs.* 42, 81 (1954).
7. RICHMAN, D. Colour-Carrier Reference Phase Synchronization Accuracy in N.T.S.C. Colour Television. *Proc. Inst. Radio Engrs.* 42, 106 (1954).
8. RICHMAN, D. The d.c. Quadri-correlator: A Two-Mode Synchronization System. *Proc. Inst. Radio Engrs.* 42, 288 (1954).
9. FARREN, L. I. Phase Detectors. *Wireless Eng.* 23, 330 (1946).
10. CLARK, E. G. A Self-Balancing Phase Detector for Colour Television Reference Oscillators. *Convention Record of the Inst. Radio Engrs.* Pt 7, 31 (1954).
11. MORRISON, W. C., KARSTAD, K., BEHREND, W. L. Test Instruments for Colour Television. *Proc. Inst. Radio Engrs.* 42, 284 (1954).
12. STERN, A. P. Phase Measurements at Subcarrier Frequency in Colour Television. *Convention Record of the Inst. Radio Engrs.* Pt 4, 57 (1953).
13. PRITCHARD, D. H., RHODES, R. N. Colour Television Signal Receiver Demodulators. *RCA Rev.* 14, 205 (1953).
14. GOOD, W. E. Colour Synchronisation in the N.T.S.C. Colour Television Receiver by Means of the Crystal Filter. *Convention Record of the Inst. Radio Engrs.* Pt 4, 9 (1953).
15. KABELL, L. R., EVANS, W. E. A Transistor Subcarrier Generator for Colour Receivers. *Trans. Inst. Radio Engrs. BTR-1*, 9 (1955).

(To be continued)

The paper-backed magnetic tape used is made in the form of a continuous loop 3½in wide, which normally lasts 30 seconds. Special loops can be supplied, however, having any duration between 10 seconds and about 4 minutes. A short length of metallic foil is spliced in at the point where the two ends of the loop join. At the end of the announcement the foil completes the circuit between two contact fingers, automatically switching off the mechanism and preparing it for the next announcement.

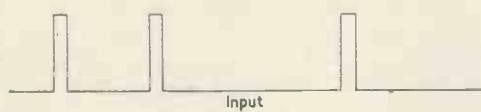
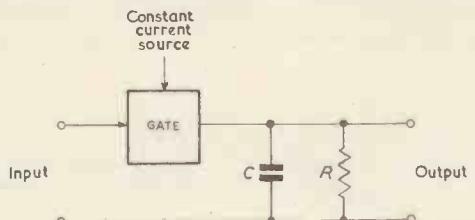
The loop of tape is moved past the playback head at a steady speed of 7½in/sec by a mechanical drive system powered by a 1/20 h.p. motor. The motor is dynamically balanced and resiliently mounted at the rear of the panel, the drive being transmitted by a flexible shaft, nylon helical gears and two large diameter pulleys coupled by a nylon loaded fabric belt. On selecting the required announcement, either by push-button or by remote control, the playback head is moved to the correct tracking position by a specially designed mechanism patented by the G.E.C. Control of the tape motor and the head positioning device is effected by a 25-position rotary selector switch associated with seven switching relays mounted under detachable covers.

# A New Serial Digital Decoder

By S. V. Soanes\*, M.A., Ph.D.

This article describes a new type of serial decoder which is virtually independent of the frequency of the clock timing pulses, unlike some previous designs. The device is fundamentally a polynomial generator, and so can also be used in analogue computers to solve certain types of polynomial equations. As nearly all digital computers use the binary system, this will be assumed to be the mode of operation throughout this article unless otherwise specified; a different radix could be used with minor changes.

In the field of digital computers there is frequent need for a device which will give an analogue output corresponding to the value of a digital number input. Such a device is called a decoder. For example, cathode-ray tube displays of output information will generally require two decoders—one each for the X and Y co-ordinates. In a parallel system, where the digits are represented by signals appearing simultaneously on separate lines, a conventional adding network, using resistors as the weighting elements, can be used as a satisfactory decoder. In a serial system,



Output

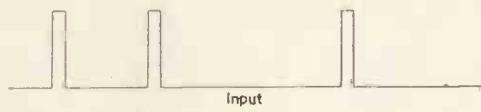
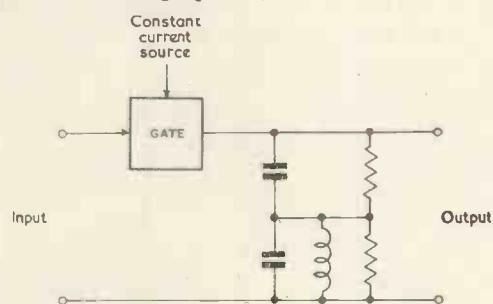
Fig. 1. Shannon decoder

all the digits of a given number appear on a single line, but in some predetermined time sequence. One way of handling this situation is to convert the serial information into its parallel form through a system of gates and clock timing pulses, and then use this method of weighted addition.

Direct conversion of serial digital to analogue form requires a different approach. A very simple form of decoder is due to C. E. Shannon<sup>1</sup> of the Bell Telephone Laboratories. Fig. 1 shows an elementary version of this. The input signal gates a fixed quantity of charge on to the capacitor  $C$  for each digit present. The time-constant  $RC$  is such that the voltage on the capacitor will have decayed to exactly one-half of its value after an interval equal to the digit period. The voltage (or charge) on this capacitor one digit period after the completion of a number will be given by:

$$\begin{aligned} & 1/2[A_n \cdot (1/2)^{n-1} + A_{n-1} \cdot (1/2)^{n-2} + \dots + A_2 \cdot (1/2) + A_1] \\ & = 1/2^n[A_1 \cdot 2^{n-1} + A_2 \cdot 2^{n-2} + \dots + A_{n-1} \cdot 2 + A_n] \end{aligned}$$

The  $A$ 's are either 0 (no digit pulse) or 1 (digit pulse present).  $A_n$  is the least significant digit and must come first in the time sequence. The disadvantage of this method is that pulse timing is very critical because of the slope of the exponential. For unity voltage at  $t = 0$ ,  $dE/dt = -0.35$  after a single digit period, and  $dE/d(RC) = +0.24$ . (Unit of time used is the digit period).



Output

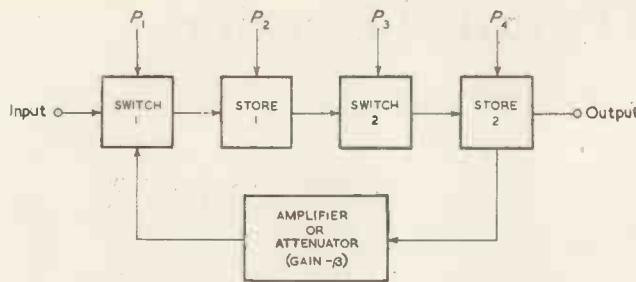
Fig. 2. Shannon-Rack decoder

An improvement by A. J. Rack<sup>1</sup> is shown in Fig. 2. This circuit can be designed to have zero first and second derivatives at the sampling times, thus reducing the errors caused by deviations in sampling times and/or component values.

In certain applications, even the Shannon-Rack decoder may be too critical as to timing. It is necessary to keep the slope of the voltage curve at zero for a longer time. This suggests the use of an intermediate storage medium with a long enough time-constant that several per cent change in the sampling time will not seriously affect the result. Greater independence on circuit values is also desirable. Furthermore in some applications it may be more convenient to have the most significant digit first in the time sequence. This is not possible with the Shannon-Rack decoder as originally described, although it would be possible with the addition of a negative resistance element.

A decoder which meets these requirements is shown in Fig. 3. The serial coded pulse input is applied to the input at switch 1. At the appropriate time, approximately coinci-

\* Ferranti Electric Limited, Toronto, Canada.



$\beta = \frac{1}{2}$  or 2 for a binary decoder

Fig. 3. Soanes decoder

dent with each input pulse, a priming pulse,  $P_1$ , is applied, which closes switch 1 and permits the input pulse, as well as the output from the amplifier or attenuator, to enter store 1 which has just been emptied by a clearing pulse,  $P_2$ . Before the next pulse of the series, a second priming pulse,  $P_3$ , transfers the contents of store 1, through switch 2, to store 2, which has just been emptied by a clearing pulse,  $P_4$ . The output of store 2 passes through an amplifier or attenuator which has a gain  $\beta$ , and this is added to the next input pulse and put into store 1 in the same way as above. The final decoded output can be sampled conveniently at the output of either store.

If the successive input pulses are  $A_1, A_2, \dots$  and the gain of the amplifier is constant at  $\beta$ , the output after  $n$  input pulses is given by:

$$A_1 \cdot \beta^{n-1} + A_2 \cdot \beta^{n-2} + \dots + A_j \cdot \beta^{n-j} + \dots + A_{n-1} \cdot \beta + A_n$$

For an  $n$ -digit serial binary decoder with the least significant digit coming first in time,  $\beta = 1/2$ , and the output is:

$$\begin{aligned} & A_1 \cdot (1/2)^{n-1} + A_2 \cdot (1/2)^{n-2} + \dots + A_{n-1} \cdot (1/2) + A_n \\ &= (1/2)^{n-1}[A_1 + A_2 \cdot 2 + \dots + A_{n-1} \cdot 2^{n-2} + A_n \cdot 2^{n-1}] \\ &= (1/2)^{n-1} \sum_{j=1}^n A_j \cdot 2^{j-1}, \text{ where } A_j = 0 \text{ or } A, \text{ a constant.} \end{aligned}$$

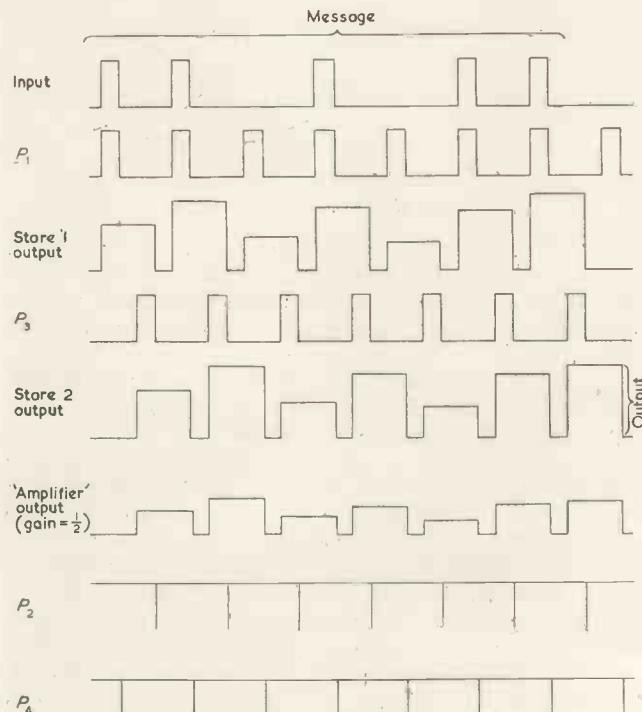


Fig. 4. Typical waveforms to decode the binary number 1101011  
(least significant digit is first)

If the most significant digit comes first,  $\beta = 2$ , and the output is:

$$\begin{aligned} & A_1 \cdot 2^{n-1} + A_2 \cdot 2^{n-2} + \dots + A_{n-1} \cdot 2 + A_n \\ &= \sum_{j=1}^n A_j \cdot 2^{n-j} \end{aligned}$$

$A$  will usually represent a voltage but can just as well be some other electrical quantity, such as current or charge, depending on the design of the system.

Typical operating waveforms for the decoding of the binary number 1101011 are shown in Fig. 4. The schematic diagram of Fig. 5 is not intended as a fully engineered circuit but is a simplified version to illustrate the principles involved.

#### Other Applications

##### (a) To DECODE DIGITAL INFORMATION IN SYSTEMS OTHER THAN BINARY (GENERAL CASE)

A decoder of radix  $r$  will require inputs of amplitudes  $A_k = 0, A, 2A, \dots, (r-1)A$ , and the gain of the amplifier must be  $1/r$ , if the least significant digit comes first in time. The output after  $n$  digits is then:

$$(1/r)^{n-1}(A_1 + A_2 \cdot r + \dots + A_{n-1} \cdot r^{n-2} + A_n \cdot r^{n-1})$$

If the most significant digit comes first in time, the gain of the amplifier must be  $r$ , and the output after  $n$  digits is then:

$$A_1 r^{n-1} + A_2 \cdot r^{n-2} + \dots + A_{n-1} \cdot r + A_n$$

##### (b) To GENERATE POLYNOMIAL FUNCTIONS

The output component after  $(n+1)$  cycles due to an input  $A_k$  is  $A_k \beta^n$ . Therefore, any polynomial of the form

$$\sum_{k,n=0}^m A_k \beta^n$$

can be generated, for  $m$  any positive integer.

This is done by applying inputs  $A_k$  at the appropriate times. These inputs may be plus or minus from the arbitrary reference level of the decoder, which level may be represented in either analogue store by zero volts or any other convenient value. If the  $n$ 's are not all integral, the evaluation can be performed by making a mathematical substitution:

$$\beta = C^q, \text{ i.e. } \beta^n = C^{qn}$$

where  $qn$  is always an integer, to a sufficient degree of accuracy. The polynomial is now of the form  $\sum_{k,n=0}^m A_k C^{qn}$

Negative values of the variable  $\beta$  could be handled by incorporating a  $180^\circ$  phase-shift in the amplifier.

##### (c) To SOLVE POLYNOMIAL EQUATIONS

An equation of the form  $f(x) = \sum_{n=0}^m A_n x^n = 0$  can be solved by evaluating it according to paragraph (b) above. The real roots of the equation can be found by varying the parameter,  $x$ , which corresponds to the gain,  $\beta$ , of the loop, and observing the values of the parameter when the output level is the same as the zero reference level of the system.

Complex roots can be found by using two decoders  $R$  and  $I$  operating effectively in parallel as follows: (refer to Fig. 6). Note that the corresponding priming and clearing pulses for each decoder are connected together.  $A$  and  $B$  are amplifiers or attenuators with the same characteristics as  $C$  and  $D$  respectively.  $E$  is a subtractor whose output is equal to (output of  $C$ ) minus (output of  $A$ ).  $F$  is an adder whose output is equal to (output of  $D$ ) plus (output of  $B$ ).

The input at terminal  $RP$  of the first decoder  $R$  is to be

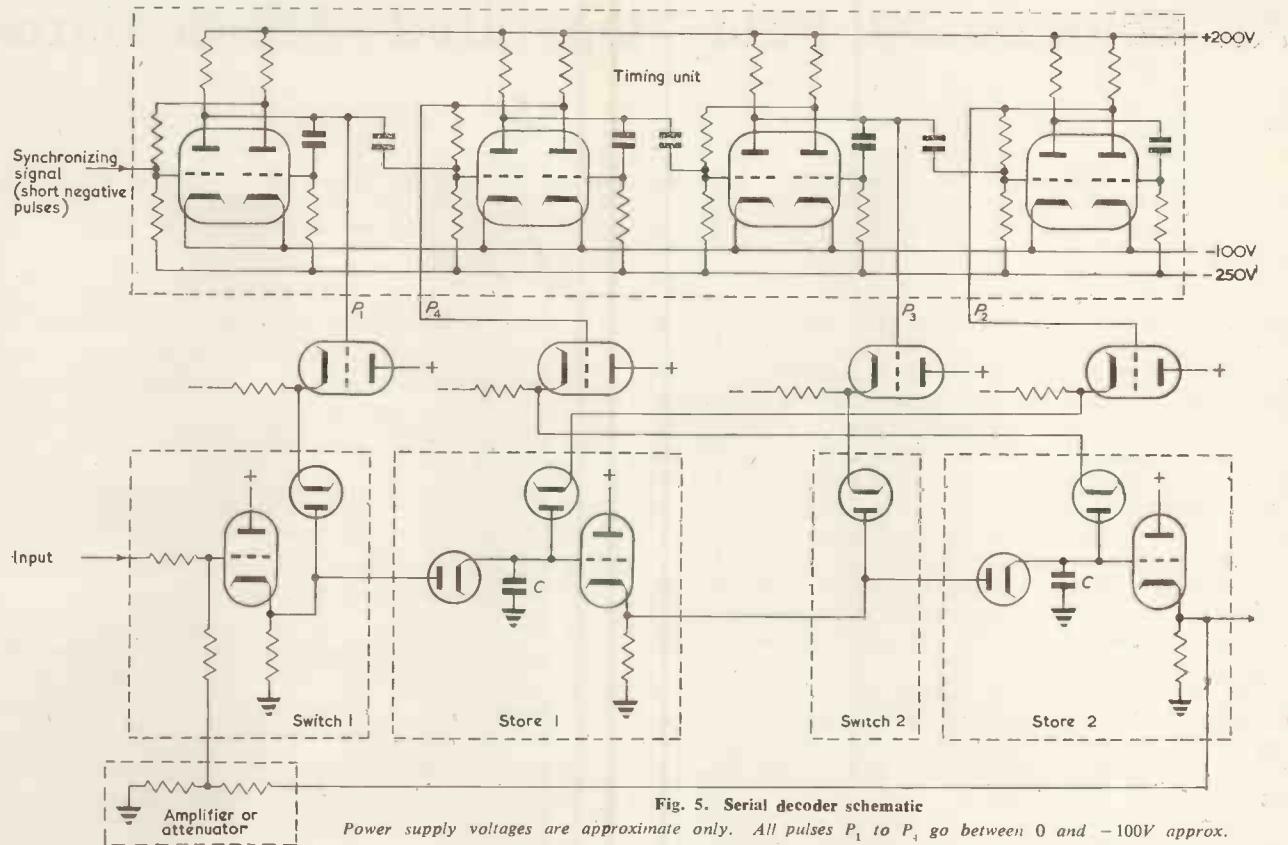
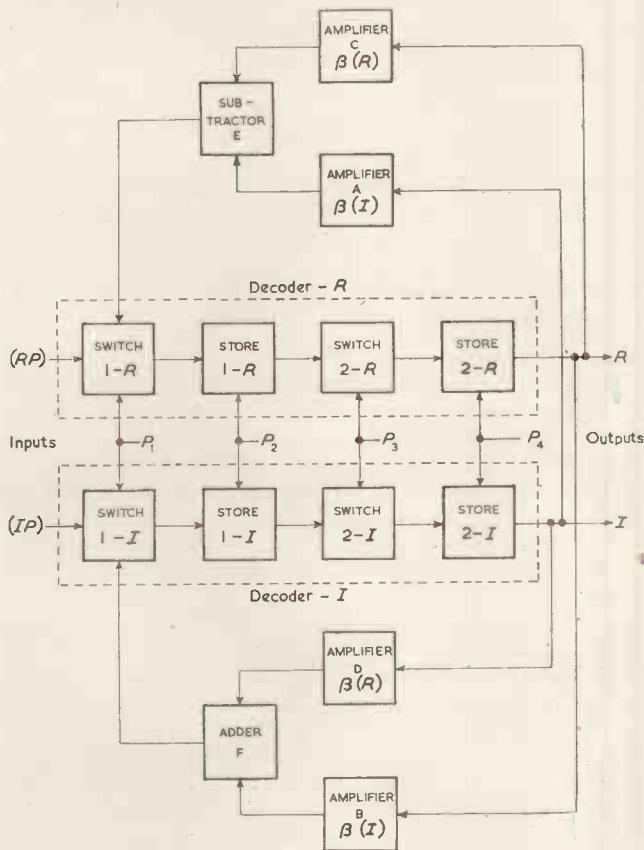


Fig. 5. Serial decoder schematic

Power supply voltages are approximate only. All pulses  $P_1$  to  $P_4$  go between 0 and  $-100V$  approx.

Fig. 6. Solution of polynomial equations with complex roots



the real part of the coefficients of the polynomial. The input at terminal  $IP$  of the second decoder  $I$  is to be the imaginary part of the coefficients of the polynomial, such that the real and imaginary parts of each coefficient are applied to the two decoders simultaneously. The output of store  $2-R$  which is applied to the input of amplifiers  $B$  and  $C$  will be designated as  $Q(R)$ . The output of store  $2-I$  which is applied to the input of amplifiers  $A$  and  $D$  will be designated as  $Q(I)$ . The roots must be of the form  $x + iy$ , and therefore two different loop gains must be provided to correspond with the two independent variables  $x$  and  $y$ , and these loop gains will be designated as  $\beta(R)$  and  $\beta(I)$  respectively. Amplifiers  $C$  and  $D$  have a gain of  $\beta(R)$ ;  $A$  and  $B$  have a gain of  $\beta(I)$ . The following equations then hold:

$$\begin{aligned} \text{(output of subtractor } E\text{)} &= Q(R) \cdot \beta(R) - Q(I) \cdot \beta(I) \\ \text{(output of adder } F\text{)} &= Q(R) \cdot \beta(I) + Q(I) \cdot \beta(R) \end{aligned}$$

$\beta(R)$  and  $\beta(I)$  are varied until the values of the parts of the polynomial, evaluated as in section (b) above, that is, the outputs of the stores  $2-R$  and  $2-I$  are simultaneously zero.

#### Conclusion

Decoders have been engineered along these principles to decode 8 binary digits with good accuracy and reliability.

#### Acknowledgments

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#### REFERENCE

- MEACHAM, L. A., PETERSON, E. An Experimental Multichannel Pulse Code Modulation System. *Bell Syst. Tech. J.* 27, 36 (1948).

# An Economical Relay Operated Accumulator

By J. K. Wood\*, M.Sc.

*An apparatus is described which sums binary numbers up to a maximum capacity of about  $2^{24}$ . The design methods are described so that a similar machine with any required capacity can be constructed. Various safety devices are described, including one to retain the information stored when the accumulator becomes full.*

RECENT work in the field of spark breakdown involved the measurement of a large number of short time intervals. These time intervals were considered in two parts; firstly, the random time-lag between the commencement of the observation and the appearance of an electron in a suitable place to initiate breakdown, usually known as the statistical time lag, and secondly the interval between the appearance of such an electron and the establishment of a conducting path across the gap, known as the formative time-lag. The conducting path is assumed to appear simultaneously with the collapse of voltage across the gap, though in certain cases (not covered in the present work) displacement current effects in the gap can invalidate this assumption.

The formative time-lag is usually assumed to be fairly constant for any one set of conditions, and an average of between 10 and 100 observations has proved quite sufficient to give a result which is consistent with similar measurements.

The statistical time-lag is quite random and may extend from about a microsecond to a second. Obtaining a reasonable overall picture of the distribution of these time-lags involves a very large number of observations. The error in the average statistical lag is inversely proportional to the square root of the number of observations, and at least 100 are necessary to give useful value for this quantity.

Conventional methods of time measurement (e.g., the ballistic galvanometer, cathode-ray oscilloscope, and various analogue devices) are unsuitable because of the narrow range of measurement which is possible. The c.r.o. with an exponential time-base is better, but many observations still fall outside its range, and the processing and measurement of the photographic recordings add very considerably to the labour. The present apparatus considerably reduces the labour and time required to produce a set of results.

A binary counter chronometer was built which measured time-lags up to 1 second, to the next lowest whole number of microseconds. The output was displayed on 20 neon indicator bulbs and taken electrically to the accumulator, which summed the readings taken, and recorded the number of readings which had been summed. The accumulator was especially useful, as it saved the binary-decimal conversion of individual readings which would otherwise have been necessary. The present apparatus allows all the working time to be used in experiment. Any necessary computation on results can be carried out during a subsequent run, as automatic sequence control is included which makes continuous supervision unnecessary. All voltage sources are stabilized.

The purpose of the present article is to describe the relay operated accumulator. This was designed to meet the following specifications:

- (1) The maximum time available for a single addition was  $12\frac{1}{2}$  sec, as this was the minimum time needed for

the sparking voltages to become stable between discharges.

- (2) The apparatus must have a very high input impedance, and in this case the input is in the form of the potential across the neon indicator lamps in the counter.
- (3) The input and the sum are in parallel form.
- (4) The apparatus must be reliable in operation.
- (5) The apparatus must be simple in design and require the minimum of components.
- (6) The accumulator must have a capacity of about 25 digits. It was anticipated that the sum of about 1 000 time intervals would only occasionally fill a 25-digit accumulator. A 20-digit number would be the largest possible input, and would occur very infrequently.

The input is in parallel form, and a parallel adder involves a large number of relays (at least 3 per digit, according to the design). An accumulator would require at least one more relay per digit, so that a 25-digit parallel accumulator would require at least 100 relays. The time available for addition is much more than that needed by a parallel adder, and such a circuit cannot be justified. A serial method is therefore used. Serial-parallel conversion is effected by a uniselector and a considerable economy in both relays and input circuits results.

## Circuit Description

### INPUT CIRCUIT

The signals from the anodes of the neon indicator lamps in the counter are taken to the uniselector and from the wiper arm, the selected signal is taken to one grid of a long-tailed pair. In the anode of the same side of the long-tailed pair is a relay ( $L$ ) which provides the signal for the accumulator. The other grid of the long-tailed pair is taken to a potentiometer, used to adjust the d.c. level of the cathodes. The circuit is given in Fig. 1(a).

### ADDING CIRCUIT

The adding circuit has to produce a sum digit and a carry digit from three input digits. These are the input digit from the counter ( $x$ ), the digit already in the accumulator ( $y$ ), and the carry digit from the previous addition ( $w$ ).

The possible inputs to the adding circuit are shown in Table 1.

The circuit which produces the sum digit  $S$  can be synthesized as follows. A closed-circuit is taken to represent 1 and an open-circuit 0.

A series of contacts which produce a closed circuit with input combination (b) (Table 1) but an open-circuit with all other input combinations is easily arrived at, and is given in Fig. 2(a). Similar circuits can be made for all the possible inputs. Selecting those circuits which are closed for inputs with a sum of 1 (neglecting the carry) and putting them in parallel, a circuit is obtained which will give the sum digit correctly for every possible input. This circuit is given in Fig. 2(b). This circuit is then simplified and rearranged to

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allow the use of changeover contacts. The final circuit is shown in Fig. 2(c).

An exactly similar argument is used to design the circuit which produces the carry digit, shown in Fig. 2(d). All the contacts are mounted on three relays, called  $X$ ,  $Y$ , and  $W$  after their respective inputs.

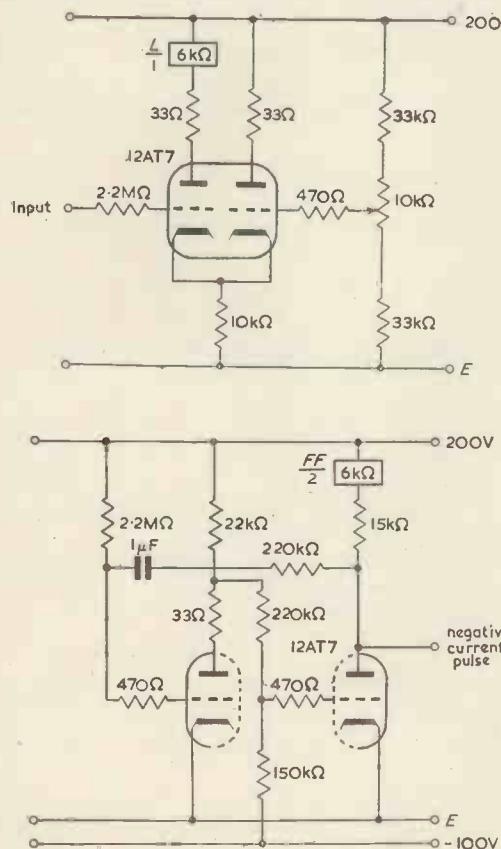


Fig. 1. Input circuits

(a) Digit input circuit to the accumulator. (b) Flip-flop; starts the addition process at the appropriate time.

TABLE 1

Input digit combinations	INPUT			OUTPUT		
	x	y	Accumulator input	w	S	W
(a)	0	0	0	0	0	0
(b)	1	0	0	0	1	0
(c)	0	1	0	0	1	0
(d)	0	0	1	1	1	0
(e)	1	1	0	0	0	1
(f)	1	0	1	1	0	1
(g)	0	1	1	1	0	1
(h)	1	1	1	1	1	1

#### THE ADDING CYCLE

The complete adding circuit has to replace the digit in the accumulator with the new sum digit, and memorize the carry digit until it is required in the next addition process. This involves a closed loop and instability is prevented by the inclusion of gating switches which control the flow of information round the loop. These gating switches are closed in turn, and are labelled  $T_1$ ,  $T_2$ ,  $T_3$ , etc., accordingly. They are open at all other times.

There are many possible cycles, but they possess features in common. A self-holding relay is needed to remember

the carry digit and other self-holding relays are needed to hold the information read from the accumulator while it is prepared to receive the new sum. Three relays will be set by the  $x$ ,  $y$ , and  $w$  inputs, and clearing circuits will be needed to clear the self-holding relays. "A" refers to the accumulator relay which is temporarily in the circuit, via the uniselector.

One method uses three input relays ( $x$ ,  $y$ , and  $w$ ) two self-holding relays, ( $S$  and  $W$ ) which hold the sum and carry digits temporarily, and a further self-holding relay

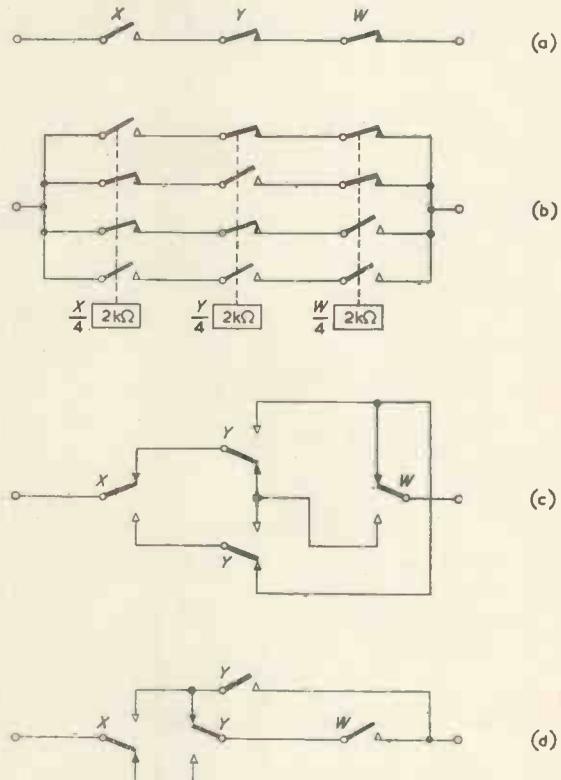


Fig. 2. Derivation of the adding circuits

(a) Contact arrangement giving closed circuit for input combination b (Table 1). (b) Contact arrangement giving the sum digit. (c) Circuit (b) simplified and rearranged to employ changeover contacts. (d) Circuit producing the carry digit.

(M) to remember the carry digit. The cycle of events is as follows:

- $T_1$ . The uniselector arrives at its next position, and relays  $x$ ,  $y$ , and  $w$  are set. The  $S$  and  $W$  relays are cleared.
- $T_2$ . The  $S$  and  $W$  relays are set from the output from the  $x$ ,  $y$ , and  $w$  adding circuits.
- $T_3$ . The inputs to the  $x$ ,  $y$ , and  $w$  relays are disconnected and the  $A$  and  $M$  relays are cleared.
- $T_4$ . The  $A$  and  $M$  relays are set from  $S$  and  $W$ .

The cycle then repeats on the next digit.

Another system makes  $x$ ,  $y$ , and  $w$  self-holding, eliminating the need for the  $S$  and  $W$  relays:

- $T_1$ . Set  $x$ ,  $y$ , and  $w$ .
- $T_2$ . Clear  $A$  and  $M$ .
- $T_3$ . Set  $A$  and  $M$  from the output from  $x$ ,  $y$ , and  $w$ .
- $T_4$ . Clear  $x$ ,  $y$ , and  $w$ . Move on the uniselector.

Alternatively, by retaining the  $S$  and  $W$  relays and making all relays self-holding, the cycle can be compressed

into three events:

- T<sub>1</sub>. Uniselector arrives at new position. Set x, y, and w; clear S and W.
- T<sub>2</sub>. Set S and W; clear A and M.
- T<sub>3</sub>. Set A from S and M from W; clear x, y, and w.

In all cases each event must last long enough to allow all relays to settle down before the beginning of the next event.

While all the above cycles are suitable for the present application, it was felt that there was scope for improvement in both speed and reliability.

It was found that the opening time of the self-holding relays was much larger than the closing time (80 to 120 msec and 10 to

30 msec respectively). Thus a cycle in which the time allowed for a clearing action is twice that allowed for a setting action, is much more efficient. Reliability of operation is improved if the number of actions which any relay undergoes is reduced. If an accumulator relay is only cleared when a 0 is to be put into it, the number of actions which that relay undergoes is reduced by 50 per cent.

#### OPERATION OF THE ACCUMULATOR

For satisfactory operation of the accumulator there are several criteria to bear in mind. These are:

- (1) Addition must begin as soon as there is a number available.
- (2) The process must stop as soon as the addition is complete.
- (3) The accumulator must refuse all inputs which occur

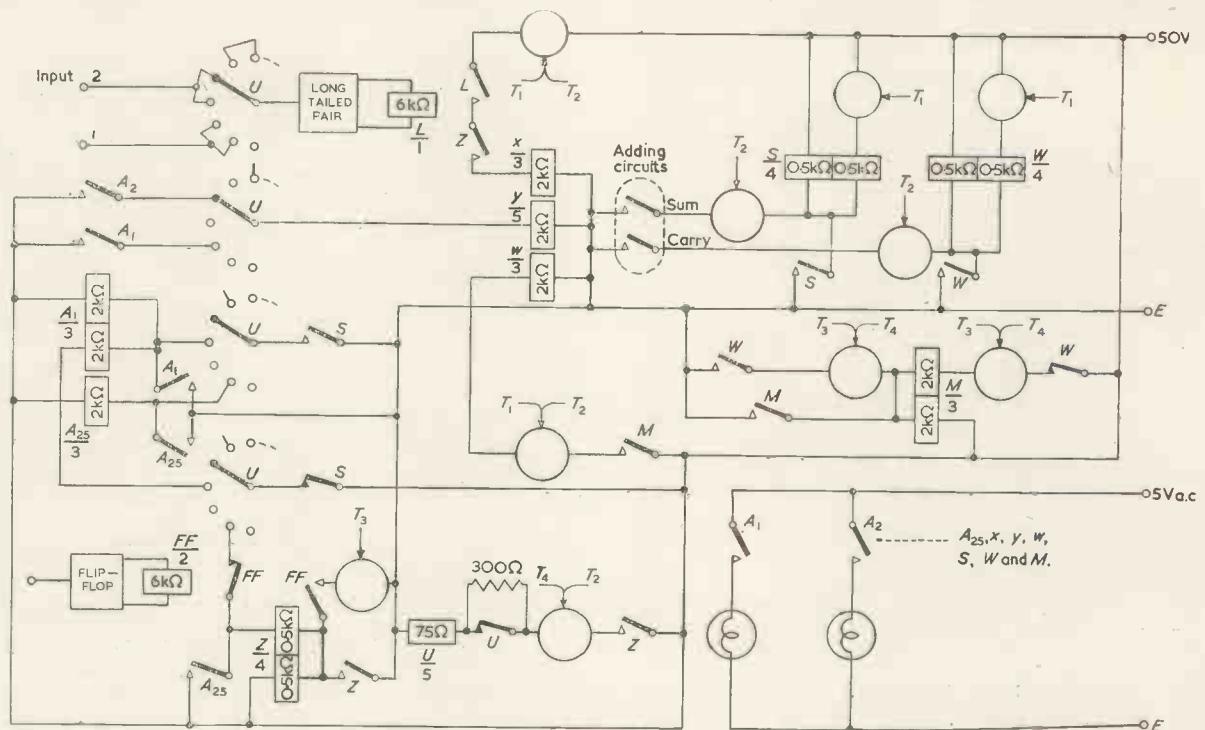


Fig. 3. The accumulator  
Accumulator relays  $A_2$  to  $A_{24}$  are exactly similar to  $A_1$  shown.

30 msec respectively). Thus a cycle in which the time allowed for a clearing action is twice that allowed for a setting action, is much more efficient. Reliability of operation is improved if the number of actions which any relay undergoes is reduced. If an accumulator relay is only cleared when a 0 is to be put into it, the number of actions which that relay undergoes is reduced by 50 per cent.

The cycle designed on these principles needs five events, but by making the S and W relays fast clearing this can be reduced to four. The majority of the self-holding relays still have twice as much time to clear as is allowed for setting. There are twenty-eight such relays.

The cycle is as follows:

- T<sub>1</sub>. Set x, y, and w; clear S and W.
- T<sub>2</sub>. Set S and W.
- T<sub>3</sub>. Adjust A from S and M from W. Disconnect x, y, and w.
- T<sub>4</sub>. Adjust A from S and M from W. Disconnect x, y, and w.

The 50 position uniselector moves on simultaneously with the change from T<sub>4</sub> to T<sub>1</sub>, and again with the change from T<sub>2</sub> to T<sub>3</sub>. The accumulator thus has a maximum capacity

of 25 digits. As alternate connexions to the uniselector are left open-circuit, the arrangement is self-gating, and correspondingly fewer contacts are required on the four-position rotary switch which controls the cycle.

- (4) The accumulator must also register how many numbers went to make up the total.

To start the addition, a high impedance pulse is received from the counter. This trips a monostable multivibrator (flip-flop) which closes relay FF connected in the anode circuit for about 1 sec, (Fig. 1(b)). Connexion is made via a gating switch (closed at T<sub>3</sub>) to set the relay marked Z in Fig. 3. This relay is self-holding. Current therefore flows through the gating switch (closed at T<sub>2</sub> and T<sub>4</sub>) to the uniselector coil and addition begins. The rotary gating switch runs continuously, and the circuit described above is necessary to ensure that the first pulse received by the uniselector coincides with event T<sub>4</sub> in the cycle.

The addition cycle is repeated twenty-five times. As the 25<sup>th</sup> digit is most significant, it only becomes 1 when the accumulator is half full or over. Before this, the clearing signal supplied by S is used to clear Z. When the accumulator becomes half full or over, no clearing pulse is supplied by S to open Z. On the other hand the 25<sup>th</sup> digit accumulator relay (A<sub>25</sub> in Fig. 3) receives a setting pulse, and is arranged to provide a clearing signal

to  $Z$ . This is made fast clearing so that the supply to the uniselector is disconnected before  $T_4$  begins, leaving the circuit ready to start the next addition.

The self-holding relay circuit is such that a clearing signal takes preference over a setting signal. Thus the accumulator is immobilized when  $A_{2s}$  supplies the clearing signal to  $Z$ , and no more additions can take place until the machine is reset to zero. A panel lamp lights up if the accumulator becomes full. This immobilization does not occur when  $Z$  is cleared from the  $S$  output, as this clearing signal is taken via contacts on  $FF$  which are opened by the pulse which starts the addition. An electromagnetic counter is triggered from  $Z$ , and used to record the total number of additions.

#### Practical Details

The self-holding relay circuit had to be arranged so that its condition was unchanged when any of the connexions to it were open-circuited. Thus both the set and clear signals must be generated by closing a circuit. Each relay has two  $2k\Omega$  coils. One is connected to the d.c. supply and the relay is set by making the circuit to ground. When the relay operates, it closes a pair of contacts which maintain the closed circuit to ground. The other coil is connected in opposition to the first, one connexion being made to the earthy side of the first coil. The free end of the second coil is connected to the d.c. supply when it is required to clear the relay. The two coils on the relay are wound to an equal resistance, so that the inner one has more turns than the outer. The inner coil is made the clearing coil, so that the resultant flux when both coils conduct is in opposition to the original flux, and therefore the remanent flux of the core material. The break is clean, though much longer than the setting time. When the relay opens the current in both coils is interrupted simultaneously by the common contact, and chatter is eliminated.

The fast-clearing relays are made from standard relays by using lower resistance coils and by increasing the air-gap in the closed position. Post Office type 3000 relays are used throughout the machine, which operates with a

basic time unit of 0.1sec. A complete addition therefore takes 10sec.

Although the capacity of the instrument is nominally 25 digits, it is actually only about 24 as the accumulator registers as full when the 25<sup>th</sup> digit becomes a 1. The sum is displayed on a control panel by a bank of indicator lamps, each fed via a pair of contacts on an accumulator relay  $A$ . These lamps are under-run to increase their life. Provision is made for testing the bulbs *in situ*, by connecting them to the supply through an independent channel. This provides a check on the bulbs without affecting the state of the machine when a reading of the sum is taken.

The accumulator is reset to zero by momentarily interrupting the 50V d.c. supply. Any spurious interruptions may cause the machine to reset, and if this is not observed, a false result will be obtained. In the present circumstances this is unlikely, as a battery supply is used. It is useful, as a check, to build into the circuit an independent self-holding fast-clearing relay and indicating lamp. If this is set at the beginning of an operation, and is still set at the end, then no interruption of supply which would affect the accumulator has taken place.

The current required to operate the uniselector is much greater than that required to maintain it in a closed state. In order to economize the power used, and to reduce the dissipation of the uniselector coils, a resistance is placed in parallel with the make and break contacts normally used for free-running. This reduces the current drain to less than one-fifth of its starting value when the contacts are opened (see Fig. 3).

The whole apparatus operates on a 50V d.c. supply, with a.c. mains for the motor driving the rotary switch and the transformer supplying the indicator lamps and the valve heaters. H.T. for the valve circuits is taken from the counter chronometer.

#### Acknowledgments

The author wishes to express his thanks to Professor F. C. Williams for facilities provided, and to the Department of Scientific and Industrial Research for a maintenance grant.

## Crane Collision Protection

At the recently held Mechanical Handling Exhibition a working model of an electrically operated overhead travelling crane demonstrated a new device developed by BTH—known as "Collimit"—for preventing collision between two overhead cranes working on the same track. Moreover, the system gives protection when cranes are working in the same bay at different levels to prevent collision between the lower crane and the load or hook of the upper crane.

Only two additional "down-shop" leads and a supply transformer are required, in addition to an electronic relay in each crane.

One of the down-shop leads is a potentiometer wire of negligible resistance/temperature coefficient material. This carries an alternating current supplied by a double-wound transformer located at the supply end. The other lead is a standard copper conductor carrying negligible current and is known as the register wire. The cranes are provided with a slipper collector for each wire.

The relay unit on the crane incorporates a potential transformer, the primary of which is connected between the register and potential collectors so that, in effect, the

primaries of the two transformers are in series, via the register wire, across the tapping points on the potentiometer; consequently the same current flows through both primaries. On each crane, the secondary of the potential transformer provides grid bias for an electronic valve controlling the limit relay. The valve is biased to be normally non-conducting, but the signal voltage—proportional to the distance between the two cranes—allows the valve to pass current to close the relay and so enable the crane to travel. As the cranes approach one another, the signal voltage falls and at a pre-determined value will cause the relay to open and prevent any further movement towards collision.

In any contingency the system will always fail to safety. For instance, if the supply to the potentiometer wire breaks down, or if contact is lost by either crane on either wire, the control valve is returned to its non-conducting state and opens the relay, so preventing any further movement of the cranes. Moreover, failure of the supply to either relay trips the appropriate travelling motion.

The "Collimit" system can be applied to three cranes if required without any additional "down-shop" leads, but another register wire is required for each additional two cranes thereafter.

# The Design of Low Frequency High-Pass RC Filters

By D. D. Crombie\*

*The theory of a high-pass RC filter using a twin-T network is discussed. The twin-T transfer characteristics are modified by including it in a feedback amplifier having suitable phase shift. The choice of compensating circuits for flattening the pass-band is also discussed, and an approximate optimum design procedure for the filter is given.*

AT frequencies less than about 150c/s the construction of  $L$  and  $C$  filters having sharp cut-off becomes difficult because of the low  $Q$ 's which are obtainable. Thiessen<sup>1</sup> has suggested the use of a twin-T rejection network, the transmission characteristics of which are altered by making it part of the feedback network of an amplifier and introducing suitable phase shifts. However, no design information relating the resultant frequency response to the introduced phase shifts was given. It is the purpose of this note to supply this information, and to indicate how suitable compensating sections may be designed for a flat pass-band response.

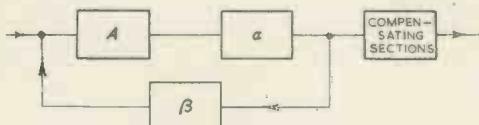


Fig. 1. Feedback network

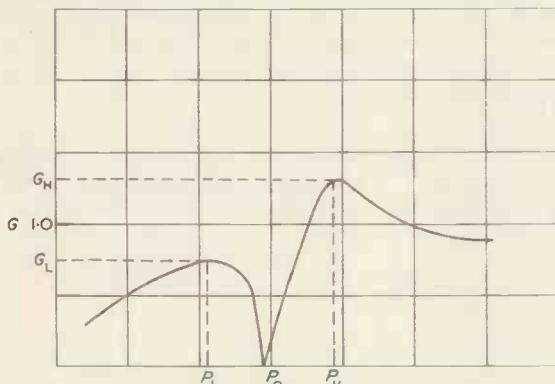


Fig. 2. Typical response curve without compensation sections

## Theory

Consider the feedback network shown in Fig. 1.  $A$  is an amplifier of gain  $A$  having a phase shift of  $\pi$ , while  $\alpha$  is a network common to the output and feedback circuits.  $\beta$  is the feedback circuit, assumed to have amplitude and phase characteristics independent of frequency. From elementary feedback theory it can be shown that the overall gain:

$$G = (e_0/e_1) = \frac{A\alpha}{1 + A\alpha\beta}$$

When  $\beta = 1$

$$G = \frac{A\alpha}{1 + A\alpha} \quad \dots \dots \dots (1)$$

Thus if  $\alpha$  has zero transmission at a particular frequency,  $G = 0$  and a rejection amplifier results. Furthermore, it can be shown<sup>1,2</sup> that incorporating the  $\beta$  network in the amplifier sharpens the null provided by the network.

It can be shown that the frequency response of a twin-T

network is of the form:

$$\alpha_1 = \frac{1}{1 - j \frac{4\rho}{\rho^2 - 1}}$$

where  $\rho = (\omega/\omega_0)$ ,  $\omega/2\pi$  being the frequency of the applied signal and  $\omega_0/2\pi$  the rejection frequency of the twin-T. The frequency response of a high-pass  $CR$  circuit can also be written as:

$$\alpha_2 = \frac{1}{1 - (j/\rho_1)}$$

where  $\rho_1 = (\omega/\omega_1)$  and  $\omega_1 = (1/C_1 R_1)$ ,  $C_1$  and  $R_1$  being the components of the  $\alpha_2$  network.

Thus if the  $\alpha$  network consists of a twin-T connected in series with a high-pass  $CR$  section, the transmission of the pair, is:

$$\alpha = \frac{1}{1 - j \frac{4\rho}{\rho^2 - 1}} \cdot \frac{1}{1 - (j/\rho_1)} \quad \dots \dots \dots (2)$$

Substitution of equation (2) in equation (1) gives:

$$|G| = A \sqrt{\left[ \left[ A + 1 - (4\omega_1/\omega_0) \cdot \frac{1}{\rho^2 - 1} \right]^2 + \left[ (\omega_1/\omega_0) (1/\rho) + \frac{4\rho}{\rho^2 - 1} \right]^2 \right]} \quad \dots \dots \dots (3)$$

giving the gain of the complete circuit.

When  $\omega_1 = 0$  this expression reduces to that for a rejection amplifier.

The expression (3) differs from that for a rejection amplifier in that when  $\rho < 1$  the gain is less than that for  $\rho > 1$  by an amount depending on  $A$  and  $\omega_1/\omega_0$ . Starting at very low frequencies the gain is approximately zero and rises before falling to zero at the null frequency. The gain then rises very rapidly when  $\omega > \omega_0$ , and then falls to a value  $A/(A + 1)$  independent of frequency as shown in Fig. 2.

The peak which occurs for values of  $\rho > 1$  is partly removed by the use of high-pass  $CR$  filters outside the feedback loop. These also reduce the subsidiary maxima which occurs below  $\omega_0$ .

## Effect of Circuit Parameters on Frequency Response

The position of the two maxima can be obtained by differentiation with respect to  $\rho$  of expression (3) and equating to zero. This leads to an equation for  $\rho$  containing powers up to 6<sup>th</sup> which is not very convenient. Approximate but more convenient results are obtained in the following manner.

Considering the high frequency maxima (at  $\rho \equiv \rho_H > 1$ ) first, it will be seen that the maximum value of  $|G|$  occurs

$$\text{when } A + 1 \approx \frac{4(\omega_1/\omega_0)}{\rho_H^2 - 1}$$

(This is true to better than one per cent for the worst conditions,  $A = 50$ ,  $(\omega_1/\omega_0) = 5$ ).

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Thus:

$$\rho_H = \sqrt{1 + \frac{4(\omega_1/\omega_0)}{A+1}} \quad \dots \dots \dots (4)$$

Similarly, the gain  $|G_H|$  for  $\rho = \rho_H$  is given by:

$$|G_H| = \frac{A}{(\omega_1/\omega_0) \cdot (1/\rho_H) + \frac{4\rho_H}{\rho_H^2 - 1}} \quad \dots \dots \dots (5)$$

Note that  $|G_H|$  never becomes infinite and thus the circuit is inherently stable.

It will be seen from equations (4) and (5) that for a fixed gain  $A$ ,  $\rho_H$  increases with  $\omega_1/\omega_0$  and that  $G_H$  passes through a maximum as  $\omega_1/\omega_0$  is increased. Similarly with  $\omega_1/\omega_0$  fixed,  $\rho_H$  decreases and  $|G_H|$  increases as  $A$  increases. Values of  $\rho_H$  and  $(G_H)$  are shown in Fig. 3 for various values of  $A$  and  $\omega_1/\omega_0$ .

The relative frequency  $\rho_L$  of the low frequency maxima is slightly less obvious but may be obtained approximately as follows:

The square of the denominator of equation (3) can be written approximately as:

$$D^2 \approx [(A+1) + (4\omega_1/\omega_0)(1 + \rho^2 + \dots)]^2 + [\omega_1/\omega_0 \cdot 1/\rho]^2$$

if  $\rho \ll 1$  and  $\omega \ll (\omega_1/4)$ .

$$\therefore D^2 \approx (A+1)^2 + (8\omega_1/\omega_0)(A+1) \cdot (1+\rho^2) + 16(\omega_1/\omega_0)^2 \cdot (1+2\rho^2) + (\omega_1/\omega_0)^2(1/\rho^2) \dots (6)$$

Differentiating:

$$(\partial D^2 / \partial \rho)_{\omega_0=\text{const}} = [8(A+1) + 32(\omega_1/\omega_0) - (\omega_1/\omega_0)(1/\rho^4)](2\omega_1/\omega_0)\rho$$

On equating to zero:

$$\rho_L = \sqrt{\left[ \frac{\omega_1/\omega_0}{8(A+1) + 32(\omega_1/\omega_0)} \right]^{1/2}} = \sqrt{\frac{\rho_H^2 - 1}{2\rho_H^2}} \quad \dots \dots \dots (7)$$

since  $\frac{4\omega_1/\omega_0}{A+1} = \rho_H^2 - 1$  from equation (4).

The corresponding gain  $|G_L|$  may be found by substituting  $\rho_L$  in equation (3). Curves for  $\rho_L$  and  $|G_L|$  are given in Fig. 4 for various values of  $A$  and  $\omega_1/\omega_0$ . The variation in  $\rho_L$  is small but  $\rho_L$  increases as  $\omega_1/\omega_0$  increases, and decreases

Fig. 3. Gain and frequency of the high frequency peak

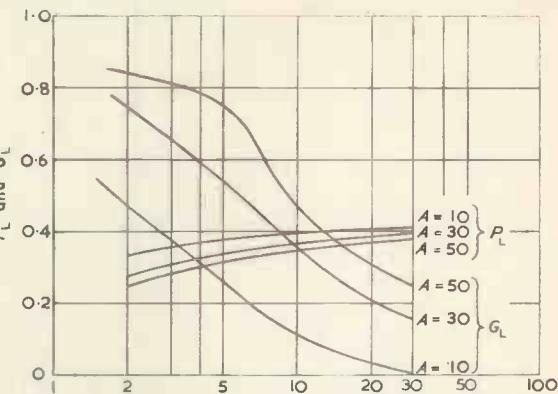
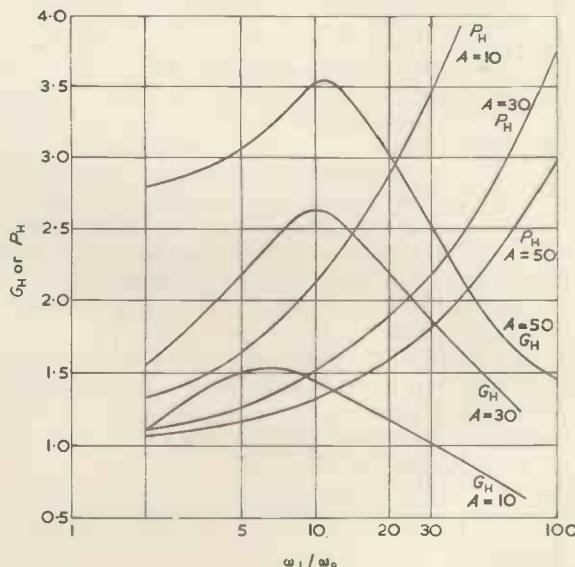


Fig. 4. Gain and frequency of low frequency peak

as  $A$  increases. However, the corresponding variations in  $|G_L|$  are much more marked:  $|G_L|$  decreasing as  $\omega_1/\omega_0$  increases, and increasing with  $A$ .

#### Compensating High-Pass Sections

As stated previously, the feedback network is in series with a  $CR$  high-pass network which is used to reduce the peak at the start of the pass-band. The presence of this section also reduces the maximum response in the stop-band. Because the peak at the edge of the pass-band is steep, the use of more than one compensating section is advisable and leads to a flatter response in the pass-band, and increased attenuation in the stop-band.

Assuming that  $n$  identical but isolated  $CR$  sections are used, then the value of  $CR$  required to reduce the peak to unity may be easily found.

The attenuation  $|G_c|$  of  $n$  such sections in series is:

$$|G_c| = \left[ \frac{1}{1 + (\omega_0/\omega)^2} \right]^{n/2} \quad \dots \dots \dots (8)$$

where  $\omega_0 = (1/R_c C_c)$  and  $\omega$  is the input angular frequency. Thus if the peak  $|G_H|$  is to be reduced to unity  $|G_c| \cdot |G_H| = 1$ .

Thus:

$$\omega_0 = \omega_H \sqrt{|G_H|^{2/n} - 1} \quad \dots \dots \dots (9)$$

The value of  $G_c$  at other frequencies is thus:

$$|G_c| = \left[ \frac{1}{1 + (\rho_H/\rho)^2 (|G_H|^{2/n} - 1)} \right]^{n/2} \quad \dots \dots \dots (10)$$

In the case of two sections  $n = 2$  and equations (9) and (10) reduce to:

$$\omega_0 = \omega_H \sqrt{|G_H| - 1} \quad \dots \dots \dots (11)$$

and:

$$|G_c| = \frac{1}{1 + (\rho_H/\rho)^2 (|G_H| - 1)} \quad \dots \dots \dots (12)$$

The effect of increasing the number of compensating sections is shown in Fig. 6 for  $n = 0, 1, 2$ . It will be seen that as  $n$  increases the filter shape improves.

#### Design Considerations

The properties usually required of a filter are that the response outside the pass-band be as small as possible and that the rate of attenuation at the edge of the pass-band be large. Thus a criterion of performance for the filter under discussion may be established and used to find the optimum parameters. To make the rate of attenuation at the edge of the pass-band as great as possible  $|G_H|/\rho_H$  should

be large. Because of the action of the compensating sections, the minimum attenuation in the stop-band will be greatest when  $(\rho_H/\rho_L)^2 \cdot |G_H|/|G_L|$  (approximately from equation (12)) is greatest. Thus we can use as a factor of merit the expression:

$$F = (|G_H|/\rho_H) \cdot (\rho_H/\rho_L)^2 (|G_H|/|G_L|) = \frac{|G_H|^2 \cdot \rho_H}{|G_L| \cdot \rho_L^2}. \quad (13)$$

This has been plotted in Fig. 5 for typical values of  $A$  and  $\omega_1/\omega_0$ . It will be seen from this that  $F$  increases with

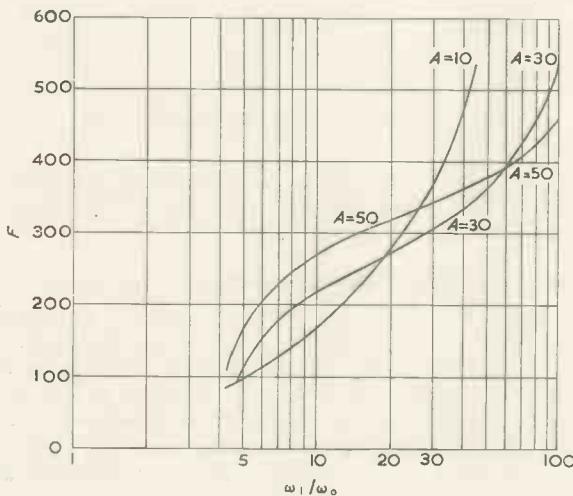


Fig. 5. Factor of merit

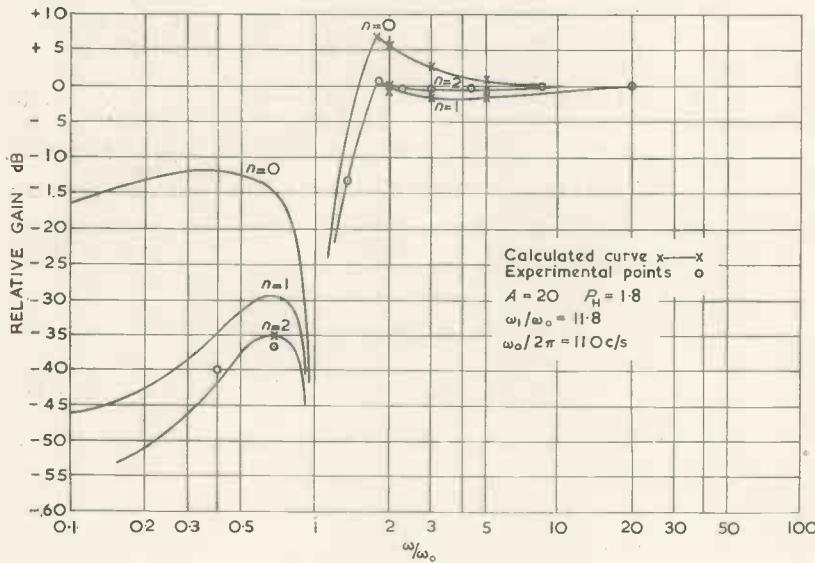


Fig. 6. Calculated and experimental results

$\omega_1/\omega_0$  and for large values of  $\omega_1/\omega_0$  is highest for low values of  $A$ . This is because  $|G_L|$  is small and  $\rho_H$  large under these conditions. For smaller values of  $\omega_1/\omega_0$  giving sharp initial cut-off ( $\rho_H$  small),  $F$  increases with  $A$ .

It will frequently happen that the values of  $\rho_H$  will be determined by the use to which the filter is put. Under these circumstances it is then necessary to pick  $A$  and  $\omega_1/\omega_0$  so that the desired value of  $\rho_H$  is obtained and so that the expression:

$$F_1 = \frac{|G_H|^2}{|G_L| \cdot \rho_L^2} \dots \dots \dots \quad (14)$$

is maximized.

Under these circumstances, it is clearly necessary to make  $A$  as large as possible to obtain large attenuation in the stop-band.

It will be seen from Fig. 6 that the effect of the compensating sections is to shift the low frequency maximum towards the null frequency and that the maximum response in the stop-band is thus greater than would be expected from multiplying the value of  $|G_L|$  and  $|G_c|$  at  $\rho_L$ . Thus the factors of merit  $F$  and  $F_1$  are only guides to the filter performance.

### An Example

A high-pass filter having a null frequency  $\omega_0/2\pi$  of 110 c/s and cut-off frequency  $\omega_H/2\pi = 200$  c/s with an amplifier

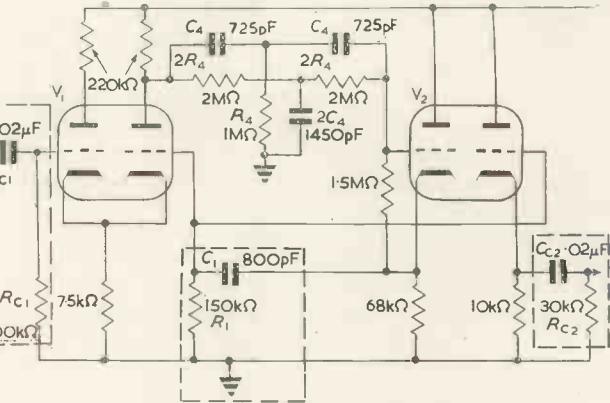


Fig. 7. Circuit diagram of high-pass filter

of gain 20 was required.

$$\text{Thus } \rho_H = 1.8 \quad A = 20$$

$$\therefore \text{from equation (4)} \quad \omega_1/\omega_0 = 11.8$$

Thus  $|G_H| = 2.22$  from equation (5) and  $\rho_L = 0.38$  while  $|G_L| = 0.24$  from equations (7) and (3).

Assuming that two compensating sections are used to make the gain at  $|G_H| = 1$ , then from equation (11):

$$\omega_0/\omega_H = \sqrt{1.22} = 1.105$$

$$\therefore \omega_0/\omega_0 = (\omega_0/\omega_H)$$

$$(\omega_H/\omega_0) = (\omega_0/\omega_H) \cdot \rho_H$$

from which  $R_c$  and  $C_c$  can be found.

The circuit is shown in Fig. 7. This circuit differs from that used by Thiessen, to avoid loading of the amplifier and twin-T circuit. The first valve  $V_1$  is used as an amplifier with the feedback applied to the grid of the second section ( $V_{1b}$ ). The first cathode-follower  $V_{2a}$  isolates the phase shifting section from the twin-T network and this avoids loading the twin-T. The second section is used to couple a low impedance load  $R_3$  without upsetting the loop gain.

The compensating sections are  $R_{c1}$ ,  $R_{c2}$ ,  $C_{c1}$ ,  $C_{c2}$ , while the phase shifting section is  $C_1R_1$ .

The calculated performance with  $n = 0, 1, 2$ , together with the measured performance for  $n = 2$ , is shown in Fig. 6. The minimum measurable signal was -40dB. It will be seen that the agreement between the theoretical curve and the measurements is quite good.

### REFERENCES

1. THIESSEN, G. J. R.C. Filter Circuits. *J. Acoust. Soc. Amer.* 16, p. 275 (1945).
2. VALLEY, G. F., WALLMAN, H. Vacuum Tube Amplifiers. Ch. 10. (McGraw-Hill, N.Y., 1948).

# Principles and Application of Electronic Analogue Computers

(Part 4)

By P. Heggs\*, A.M.I.E.E.

## A Complete Computer Installation

A front view of the Saunders-Roe analogue computer is shown in Fig. 24.

The operational amplifiers arranged in blocks of eight are situated at the top of the console on either side of the patch panel. Above the operator's desk, from left to right, are the twin pen recorder, control panel and cathode-ray oscilloscope. The output of each amplifier has been monitored by a 1½in centre reading meter engraved 1-0-1 machine units. Beneath these and symmetrically disposed on either side of the desk are forty ten-turn helical potentiometers, and below these are one hundred and eighty high quality type single-turn potentiometers, which are permanently wired for the six degrees of freedom aircraft flutter equations.

The machine operations are controlled by Kellog keys situated on the control panel, the operator in the picture being shown with his finger on the start key.

Power supplies are housed below the desk with h.t. stabilizers on each side of these units, and underneath these are the panels containing all heater supplies, fed from a mains voltage stabilizer. On the extreme right is seen an aircraft auto-stabilizer, which has been arranged to work in conjunction with a set of aircraft equations, simulated by components shown on the patch panel. The far left-hand rack contains a diode limiter panel employing six double diodes. Above the auto-stabilizer equipment is the function generator containing constant voltage sources for calibration and forcing functions. The functions generated are ramps, or linear rises,  $(1 - e^{-at})$ , triangles and impulses. These may be operated on by  $pT$  /  $(1 + pT)$  or  $1/(1 + pT)$  to form approximate derivatives or integrals. Any of the above functions may be generated once or repetitively by the operation of a switch.

## Applications of Analogue Computers

Consider the coupled mechanical system described in Part 1, which consisted of an inertia and a pendulum, coupled by a spring to a similar system. This problem has been taken to show the ease with which results may be obtained when several parameters of a set of coupled elements require to be varied at once.

Using problem variables,  $\theta$ , and the equations of motion become:

$$(m_1 r_1^2 + I_1) (d^2\phi_1/dt^2) + f_1(d\phi_1/dt) + (k + m_1 g r_1)\phi_1 - k\theta_1 = 0$$

$$(m_2 r_2^2 + I_2) (d^2\theta_1/dt^2) + f_2(d\theta_1/dt) + (k + m_2 g r_2)\theta_1 - k_1 = 0$$

Suppose that all the values in this system are fixed apart from the lengths of each pendulum, and the form of the oscillations resulting when the system is disturbed from rest is required for varying lengths  $r_1$  and  $r_2$ . Let the known constants have values as follows:

$$(2r_1^2 + 10)\phi_1'' + 2\phi_1' + (20 + 64r_1)\phi_1 - 20\theta_1 = 0$$

$$(3r_2^2 + 25)\theta_1'' + 3\theta_1' + (20 + 96r_2)\theta_1 - 20\phi_1 = 0$$

Letting the highest value of  $r_1$  and  $r_2 = 1$  unit of length, and dividing through each equation by the highest coefficient gives the machine equations.

$$(0.03r_1^2 + 0.15)\phi'' + 0.031\phi' + (0.31 + r_1)\phi - 0.31\theta = 0$$

$$(0.03r_2^2 + 0.26)\theta'' + 0.031\theta' + (0.21 + r_2)\theta - 0.21\phi = 0$$

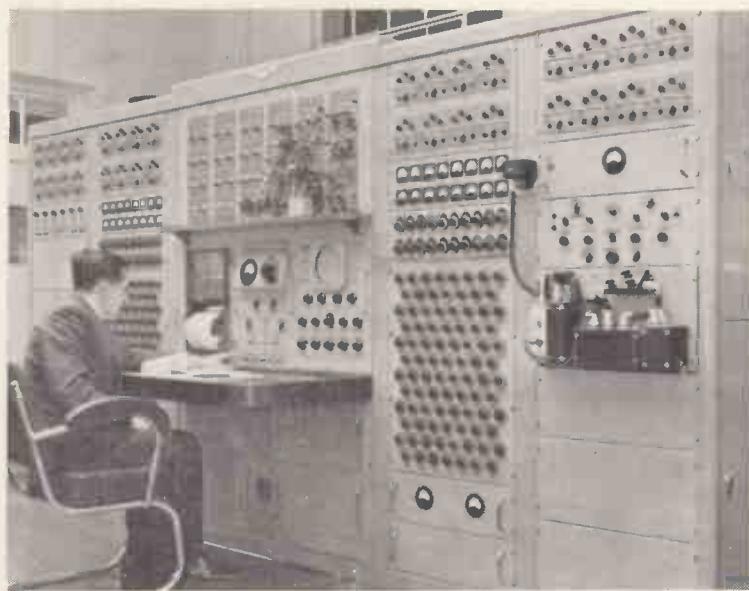


Fig. 24. The Saunders-Roe general purpose analogue computer

where  $\theta$  and  $\phi$  are now machine variables.

Now if all the integrator gains are made unity these equations may be very easily set out, and no further scaling will be required. Two separate amplifiers are required to form the terms  $0.03r_1^2\phi'$  and  $r_1$  in each channel. The computer diagram for one degree of freedom is shown in Fig. 25.

Variation in the value of  $r^2$  to a maximum of unity is obtained by the potentiometer (a) in conjunction with the variable feedback resistor (b), likewise variation in  $r$  is obtained by potentiometer (c), provided that the external gain of the amplifier generating  $r_1\phi$  is unity, and that unity gain is obtained from the amplifier generating  $r_1^2p^2\phi$  when potentiometer (a) is set at maximum.

These potentiometers may be set separately, but the procedure is tedious if many are involved, consequently if much work of this nature is undertaken ganged potentiometers are advisable.

In Fig. 25 twice the variation in the value of  $r_1$  may be obtained by halving the value of the input resistance to the  $r_1\phi$  amplifier, and dividing by four the input resistance to the  $r_1^2p^2\phi$  amplifier.

It will be noticed that two of the input potentiometers

\* Canadian Westinghouse Co. Ltd, formerly Saunders-Roe Ltd.

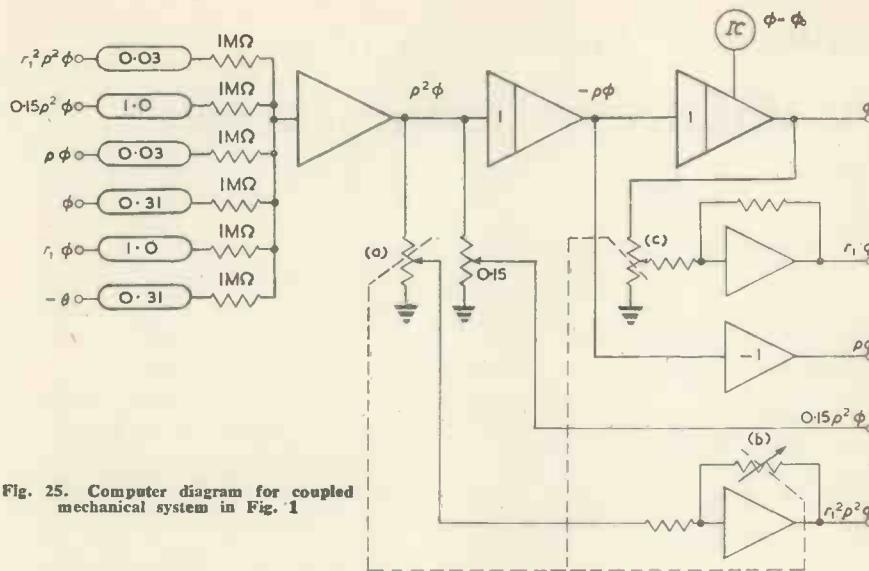


Fig. 25. Computer diagram for coupled mechanical system in Fig. 1

are set to a value of 0.03. This is well within the capabilities of a ten-turn helical potentiometer, but if a single turn component be used setting accuracy will be improved by arranging that a value of  $> 0.1$  is required.

This is best achieved by rescaling the problem by a change of integrator gains, while keeping their product constant, in order to preserve the same time scale.

Only one degree of freedom has been shown in Fig. 25, but the  $\theta$  channel is of identical form. It will be noted that coupling from the  $\theta$  equation is introduced into the  $\phi$  equation at the summation amplifier, and in like manner the  $\phi$  term is introduced into the  $\theta$  system. The system is given a disturbance as an initial condition in any integrator, or a step or impulse may be fed to one or both of the summation amplifiers. All points with the same terms present are connected together.

#### Aircraft Automatic Stabilizer

Stabilization of an aircraft in pitch is described by the following simplified equations of correct general form:

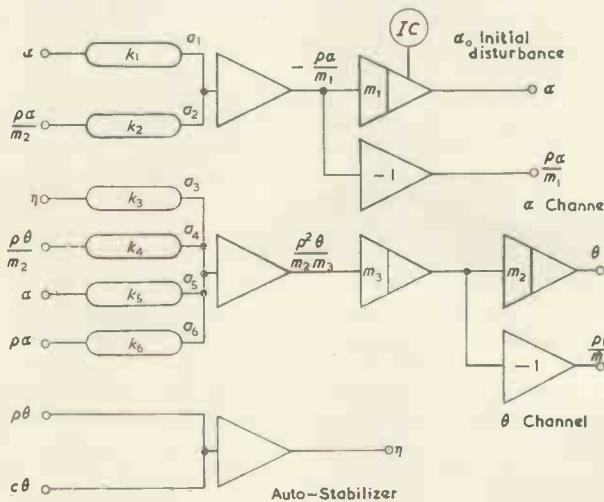
$$\alpha' = \theta' + K_1 \alpha$$

$$\theta'' = K_2 \theta' + K_3 \alpha + K_4 \eta$$

and the equation of the stabilizer is:

$$\eta = K \theta' + c \theta$$

Fig. 26. Computer diagram for aircraft equations and auto-stabilizer



where  $\alpha$  is the angle of incidence  
 $\eta$  is the elevator angle  
 $\theta$  is the angle of pitch.

The computer diagram for these equations is shown in Fig. 26, the complete problem together with the stabilizer set up on the analogue computer in Fig. 24, while Fig. 27 shows a close up view of the stabilizer.

If the  $\theta$  channel is multiplied out, by equating the output to the input of the summation amplifier and reversing the sign there obtains:

$$(p^2 \theta / m_2 m_3) = -[a_3 k_3 + (a_4 k_4 / m_2) p \theta + a_5 k_5 \alpha + a_6 k_6 p \alpha]$$

or:

$$p^2 \theta = -(a_3 k_3 m_2 m_3 \eta + a_4 k_4 m_2 p \theta + a_5 k_5 \alpha + a_6 k_6 p \alpha)$$

which compares with:

$$p^2 \theta = -(K_4 \eta + K_2 p \theta + K_3 \alpha + K_4 \eta)$$

where  $m$  = integrator gain

$k$  = potentiometer setting.

$a$  = amplifier external gain.

The computer is set up by coefficient comparison as before.

#### Solution of Partial Differential Equations

Certain partial differential equations may be solved on an analogue computer, examples of which are the equation of wave propagation:

$$(\partial^2 u / \partial x^2) = (1/C^2) (\partial^2 u / \partial t^2)$$

and the equation of heat conduction  $K(\partial^2 u / \partial x^2) = (\partial u / \partial t)$  in the latter case  $K = (k/c\rho)$ , where  $k$  = thermal conductivity,  $\rho$  = density,  $c$  = the specific heat of the medium and  $u$  = temperature.

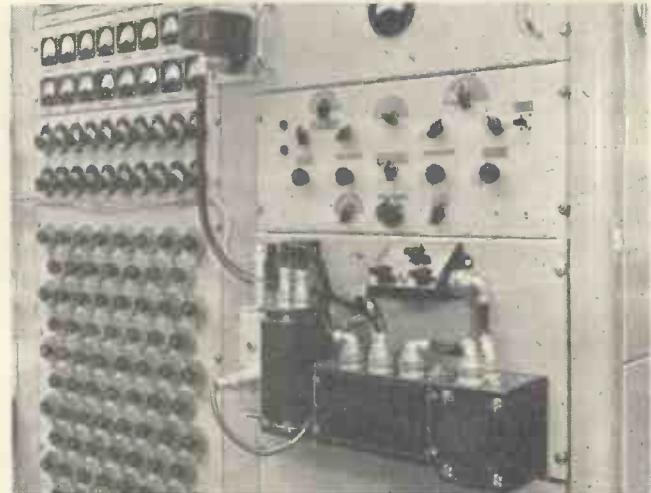
The equation is solved by expressing the term  $\partial^2 u / \partial x^2$  in finite differences.

Consider a bar of material shown in Fig. 28.

It is assumed that one end is in contact with a source of heat at temperature  $u_s$  while the latter end terminates in an insulator. The problem is one of finding the temperature at any point in the bar as a function of time.

The bar is divided up into a number of stations  $n$ .

Fig. 27. Aircraft auto-stabilizer working in conjunction with computer



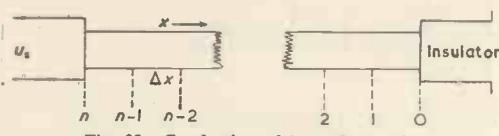


Fig. 28. Conduction of heat along a bar

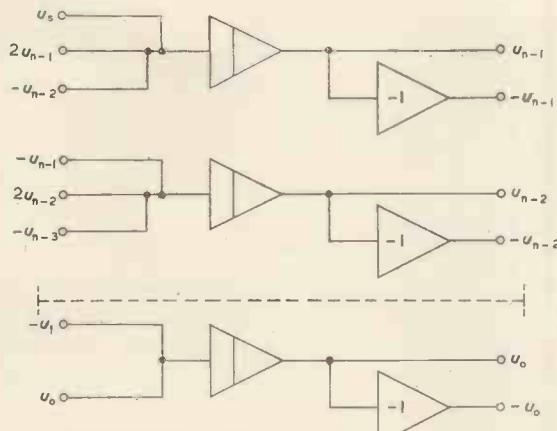


Fig. 29. Computer diagram for the heat conduction equation

$n-1, n-2 \dots 1, 0$ , separated by a distance  $\Delta x$ . Then the problem is resolved into a number of simple differential equations which apply to each station.

The rate of change of  $u$  with  $x$  at the point  $n-\frac{1}{2}$  is approximately:

$$\frac{du}{dx} = \frac{u_n - u_{n-1}}{\Delta x}$$

and for the point  $n-1$  the second derivative:

$$(d^2u/dx^2)_{n-1} = \frac{(du/dx)_{n-1/2} - (du/dx)_{n-3/2}}{\Delta x}$$

### MAGNETIC SLIP RINGS \*

The problem of transferring power or a signal from a rotating shaft to a fixed circuit, or vice versa, is a common one which has until recently been solved with the aid of mechanical slip rings and brushes. However, their use in communication circuits is unsatisfactory because they introduce noise and require constant attention to avoid or overcome the effect of wear.

A new form of slip ring, which has many advantages, has recently been developed for this purpose. The new device is a magnetic slip ring and it is most unfortunate that the abbreviation "magslip" should already have been used for another device which bears no relation to a slip ring because the degree of coupling between the fixed and moving portions is dependent upon the angle of rotation whereas in a true slip ring the coupling is constant for all angles. The magnetic slip ring consists of a transformer whose core and windings are so arranged that the path of the flux is not altered with relation to the windings when the rotor revolves, so that the reluctance of the path and the coupling between the two windings remain constant. Thus, the most effective construction is one which has circular symmetry and, in general, this must eliminate the use of stampings. Fortunately, dust cores and ferrites are often moulded in circular coaxial form and these are admirable for the purpose.

The design of the windings follows normal transformer practice and it is, therefore, unnecessary to describe it in detail. It should be noted that the magnetic slip ring has some very great advantages over the mechanical slip ring; it does not generate noise, wear and maintenance are greatly reduced and it can act also as a transformer. In fact, the only disadvantage appears to be that it must occupy more space than a conventional transformer of the same power-handling capacity and

$$= \frac{u_n - 2u_{n-1} + u_{n-2}}{(\Delta x)^2}$$

so that for this station:

$$(du/dt)_{n-1} = K/(\Delta x)^2 (u_n - 2u_{n-1} + u_{n-2}) \dots \dots \dots (19)$$

The boundary conditions are that at  $x = 0$  the temperature is  $u_s$  for all values of time.

It is assumed that no heat flows out of the bar by radiation, and that no heat can be conducted from the end of the bar against the insulator, so that for  $x = l$   $(du/dx)_{x=l} = 0$

$$\text{Thus approximately } (du/dx)_{x=1} = \frac{(u_1 - u_0)}{\Delta x} = 0$$

(true for  $\Delta x \rightarrow 0$ )

$$\text{so that for } n = 0 \quad (du/dt)_{n=0} = \frac{K(u_1 - u_0)}{(\Delta x)^2}$$

The computer diagram is built up from equation (19) using one integrator per station and scaled so that  $K/(\Delta x)^2 = 1$  shown in Fig. 29.

#### Acknowledgments

The author is indebted to the Directors of Saunders-Roe Limited for permission to publish this article, to Mr. P. G. S. Jackson for suggestions in the application of Analogue Computers and to Messrs. A. Buchan, A. B. Johnson, D. P. Hames, H. R. Blackwell, whose efforts were a major contribution to the success of the Saunders-Roe project.

#### BIBLIOGRAPHY

1. KORN, G. A., KORN, T. M. *Electronic Analogue Computers*. (McGraw-Hill, 1952.)
2. Proceedings of the Western Computer Conference.
3. MEDKEFF, —, MATTHEWS, —, Solving Process Control Problems by Analogue Computer. *Instruments and Automation* (1954).
4. VALLEY, G. E., WALLMAN, H. Vacuum Tube Amplifiers. (McGraw-Hill, 1948.)
5. GREENWOOD, L. A., HOLDOM, J. V., MACRAE, D. Electronic Instruments. (McGraw-Hill.)

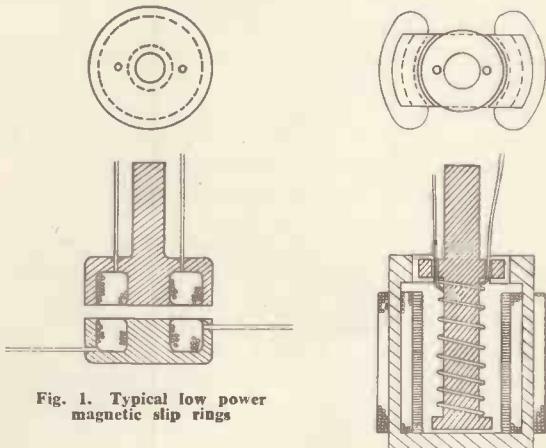


Fig. 1. Typical low power magnetic slip rings

it may be bigger than the equivalent mechanical slip rings. The increase in size is due to the fact that the dust or ferrite core has lower permeability and smaller saturation value of flux density than those of laminated core materials. The length of the gap can be made adjustable, if desired, but it should, in any case, not be too short, since this will magnify the effect of any surface irregularities and if it is extremely short it can even result in wear.

Magnetic slip rings can be made to transfer power at high as well as at low frequencies. If the two halves of the transformer are each embedded in a suitable resin, the device can be made weather-proof and extremely reliable.

Fig. 1 shows two typical samples of low power magnetic slip rings.

\* Communication from E.M.I. Engineering Development Ltd.

# An A.C. Voltage Stabilizer

By F. A. Benson\*, M.Eng., Ph.D., A.M.I.E.E., M.I.R.E., and M. S. Seaman\*, B.Eng.

In this mains-voltage stabilizer a temperature-limited diode is used as the controlling element. The diode is placed in one arm of a bridge circuit, the other three arms containing resistors. The bridge is balanced at a particular value of diode-filament voltage, but a change in the latter gives an unbalance voltage which is amplified and is then used to control the voltage requiring stabilization via a saturable-core reactor with an associated auto-transformer. A brief review of previous work on several types of saturated diode suitable for this purpose is given before the stabilizer and its performance are described.

THERE are many electronic applications in industry, communications work and in research laboratories where a high degree of constancy of the mains voltage is required. A demand therefore exists for a voltage stabilizer which is capable of supplying a variable load irrespective of variations in both mains voltage and frequency. The equipment should preferably deliver an output voltage with a good waveform, but if distortion is present then the r.m.s. value of the voltage should remain constant.

Many types of a.c. stabilizers have been marketed and also described in the literature, but they suffer from several defects. Constant-voltage transformers have been produced, but they are generally frequency dependent, and often also give an inferior performance with reactive loads. If compensation for frequency variation is provided, this generally entails using a fixed load. Patchett's precision units<sup>1</sup>, while having a very good performance, have only a limited range of input voltage and are rather complicated, besides requiring special protective circuits. Richards's<sup>2</sup> and Thomas's<sup>3</sup> circuits appear to have the disadvantage of inferior performance with reactive loads. Various mechanical units have also been developed, two such circuits having been described by Long<sup>4</sup> and by Collinge and Marsham<sup>5</sup>. These schemes are characterized by their slow response time, the intermittent nature of their operation and the need for regular maintenance.

The electronically-controlled transductor circuit described here was developed for ordinary laboratory use. The basic circuit adopted, now fairly well known in principle, is similar to that described by Helterline<sup>6,7</sup> in which a temperature-limited diode is used as the controlling element.

If a diode valve is operated with a sufficiently high anode voltage, the anode current is nearly independent of the anode voltage, but depends very much on the filament voltage. The diode is placed in one arm of a bridge circuit, the other arms being composed of resistors. This bridge will be balanced at a certain filament voltage, but a change of filament voltage of a fraction of one per cent will give an unbalance voltage of several volts. This unbalance voltage is applied to the grid of a pentode or beam tetrode, the output of which is used to control the voltage requiring regulation via a saturable core reactor with an associated auto-transformer.

The saturated diode has been found to be the most suitable voltage-sensitive element which responds to the r.m.s. value, the other alternatives (lamps, barretters and thermistors) being unduly sensitive to ambient temperature and having rather large power consumption or other circuit difficulties. The power consumption of the diode and its associated circuits in this stabilizer amounts to some twelve watts.

## Saturated-Diode Characteristics

A special diode, type 29C1, has been produced for use in such circuits. The characteristics of a number of these diodes have been studied by Attree<sup>8</sup> and the authors<sup>9</sup>, typical characteristics being shown in Fig. 1. The relation

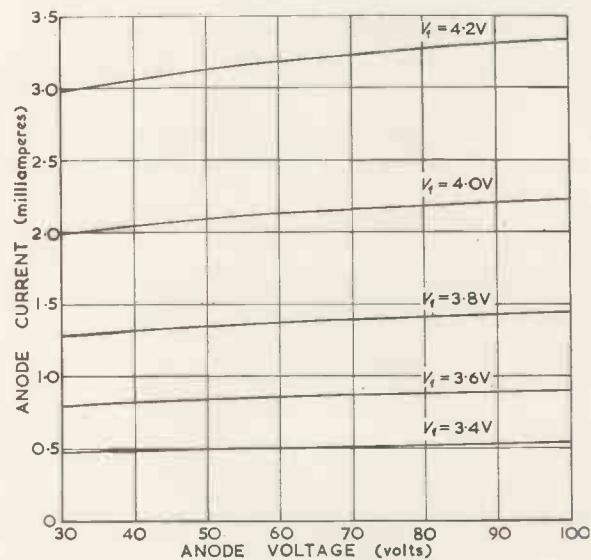


Fig. 1. 29C1 characteristics

between the emission current  $I_a$  and the filament voltage  $V_f$  is expressed by an equation of the type  $I_a = K \cdot V_f^n$ , where  $K$  and  $n$  are constants. Attree finds that with an anode voltage of 50V  $n$  is 8.5 at a filament voltage of 4.0V, so that a 1 per cent change in filament voltage would cause a change of 8.5 per cent in anode current. The authors have found that over the range of anode voltage of 30 to 100V,  $n$  is almost independent of anode voltage, values between 9.8 and 8.4 having been obtained over the filament voltage range 3.0 to 4.2V.

The warming-up period of the diode is not very long, and for all practical purposes it is virtually complete in about 5 minutes and the emission current is unaffected by changes in ambient temperature. The characteristics can sometimes be altered by as much as 0.2 per cent by a sharp blow on the glass envelope, although the initial conditions can be restored by switching off the filament for a few seconds.

The variations between various diodes have also been investigated, and although all valves have similar charac-

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teristics, quite large variations—up to  $\pm 15$  per cent—may be found in the anode current at given electrode voltages. During life, the anode current at given voltages gradually decreases with the time, although small variations of a random nature occur. The rate of decrease depends on the filament voltage, being less at lower voltages. The authors have found that the current decreases by about 10 per cent in the first 250 hours and 25 per cent in 1500 hours at a constant filament voltage of 4.0V, compared with 10 per cent in 500 hours and 25 per cent in 2100 hours at a filament voltage of 3.5V, the anode voltage being 90V in both cases, the filament supply being direct current.

Attrie<sup>8</sup> has carried out tests with constant emission currents of 5mA and 0.5mA, the filament supply being alternating current. Although he points out that the life characteristics with a.c. heating will not necessarily be the same as with d.c. heating, the two sets of results agree fairly well. Attrie bases this assumption on effects observed by Johnson<sup>10</sup> in lamps and valves with tungsten filaments, where it has been found that filaments heated by a.c. remain uniformly smooth on the surface, while those heated by direct current develop a step-like surface structure. Johnson suggests that surface migration of tungsten ions is responsible for these effects.

Hall<sup>11</sup> has made some comments on the article by the authors, based on work carried out during the development of the 29C1 valve, in which he says that the characteristics are a compromise between economical production and uniformity of performance.

The characteristics of several other valves, namely the GRD6 (now superseded by the GRD7), A2087 and the Australian type AV33 have also been investigated by the authors for use in stabilizer circuits. The investigations have been concerned with both static and fairly long-term tests. From the results of these tests, the 29C1 has been selected for the present purpose, although the AV33 is equally suitable, and can, in fact, be used in the circuit to be described without any modifications. The GRD6 has a relatively large filament which consumes rather a lot of power, apart from drifting which is troublesome. The valve envelope also gets fairly hot, and according to Attrie, its life characteristic is not so good. Its response time would appear to be rather large. The A2087 also has a very poor life characteristic. Although it has a lower filament consumption than the 29C1, its envelope also gets rather warm in places, due mainly to its small size. Its response time is somewhat longer than the 29C1 when operated under equivalent conditions.

Both the GRD6 and A2087 valves have the further disadvantage that they have a single filament wire, compared with two wires in parallel in both the 29C1 and AV33. The advantage of the latter arrangement arises when the filament eventually fails. With the GRD6 or A2087, the grid of the amplifier valve in this present circuit would then be driven positive, thus saturating the reactor and causing the output voltage to rise to a very high value. With the 29C1 or AV33, the probability of both filament wires breaking simultaneously is fairly remote, so that a failure of one wire would result in a rise in output voltage until the same total emission current is obtained from the remaining wire. This rise would be of the order of 10 per cent to 15 per cent, which should not cause damage to most loads before this partial failure was noticed and steps taken to correct matters.

In general saturated diodes are more sensitive to changes in filament current than in filament voltage, a change of 1 per cent in filament current giving a change of about 14 per cent in anode current. This effect is due to the non-

linear relationship between filament voltage and current. Such "constant-current" operation can be achieved by inserting a large resistance in series with the filament. As the filament current is of the order of 0.8A, such operation will be wasteful of power. However, to allow for ageing of the valves and also for variations between valves, some resistance in the filament circuit is necessary. If the voltage across the resistor is about half that across the filament, the sensitivity of the diode can be increased without undue losses. According to Attrie valves operated in such circuits also exhibit a better long-term stability than those operated with very little resistance in the filament circuit.

The 29C1 diode can be operated over quite a range of filament voltage, the value chosen being a compromise between several conflicting requirements. Keeping the filament voltage low gives better long-term stability, but the response time will be longer. In addition, when operated on an a.c. filament supply, there is an alternating current at twice supply frequency superimposed on the direct emission current. The relative magnitude of this a.c. component is greater at higher filament voltages, and is nearly independent of anode voltage, providing that the valve is operating in the saturated region throughout the cycle. This alternating current can be troublesome in later stages of the stabilizer circuit.

It has also been found preferable to use the diode in a 1:1 ratio bridge circuit. The sensitivity of the bridge circuit is theoretically proportional to the voltage across the anode load, so that it would appear preferable to make this as large as possible. Experimental results have shown that this is nearly true, but when the voltage across the load is not nominally the same as that on the diode anode, drifting troubles will be experienced. This has been found to be due to changes in the anode dissipation when the filament voltage changes. With a 1:1 bridge, there is very little change in anode dissipation with changes in filament voltage, and drifting troubles have been negligible. When a 2:1 ratio bridge was tried in this unit, the performance was inferior and considerable drifting was evident after changes in input voltage or load, although the bridge was theoretically more sensitive.

### The Stabilizer and its Performance

The circuit of the unit is shown in Fig. 2 and its performance is illustrated by Fig. 3. The voltage-sensitive element is the saturated diode type 29C1, which controls the saturation of the reactor via the amplifier valve EF55. The reactor controls the output voltage in conjunction with the auto-transformer.

If the reactor and auto-transformer are assumed to be perfect (i.e. no losses, no magnetizing current or leakage reactance) then the operation of the correcting circuit can be understood by referring to the vector diagram in Fig. 4. With a resistive load, output current and voltage are in phase, so that to achieve an m.m.f. balance on the auto-transformer the reactor current must be in phase with the output current. As the reactor voltage leads its current by  $90^\circ$ , it will also lead the output voltage by this angle. The magnitude of the voltages on the reactor and auto-transformer are such as to complete the vector diagram, bearing in mind the magnitude of the input voltage. It will be observed that the input current will always lag behind the input voltage unless the load current is leading by a fairly large angle, when stabilization is nearly impossible due to the large voltages on the reactor and auto-transformer. With lagging loads, these voltages are reduced, and stabilization is possible with loads of very poor power factor. Since the reactor and auto-transformer

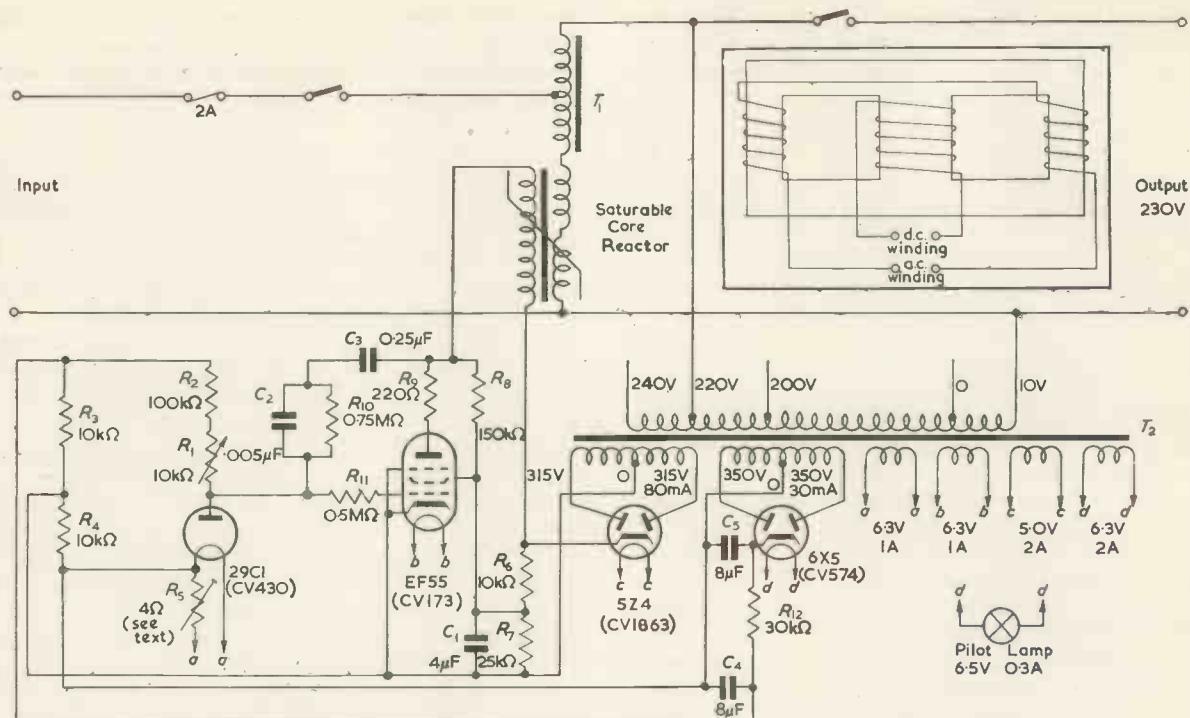


Fig. 2. Circuit of unit  
(Inset) Method of winding the saturable core reactor

are not perfect, the reactor voltage leads the output voltage by an angle somewhat less than is indicated by the above theory, and in general the magnitudes of the voltages on the auto-transformer and reactor are also less. This effect is very marked if the auto-transformer has a large leakage reactance, although under such conditions the performance falls off slightly.

The optimum ratio for the auto-transformer and number of turns on the reactor are best found experimentally. If there are too many turns on the reactor, then fairly large changes in the control current are required to counteract any disturbance, consequently stabilization and regulation are not so good; also the maximum load that the unit can supply will be less. On the other hand, if there are too few turns, the control winding cannot exercise sufficient control except over a limited range. The effect of varying the reactor turns at a fixed transformer ratio is shown in Fig. 5.

The effect of varying the auto-transformer ratio is shown in Fig. 6. Moving the input tapping towards the output end improves the power output (and the efficiency and input power factor) of the unit, but the reactor then has poor control over the output, while moving the tapping point towards the reactor end gives a more stable output with a reduction in input power. In general, the optimum reactor turns and auto-transformer ratio are not very critical.

The amplifier valve should have a high mutual conductance and a high anode slope resistance. The former requirement leads to the greatest control current change for a given input signal, while the latter means that the dynamic mutual conductance will not be much less than the static value. Thus a valve with a pentode characteristic is required, which is also capable of passing a fairly large anode current, otherwise a high supply voltage and a large number of turns of fine wire on the reactor will be required. The valve adopted was the EF55, although the 6L6 has also been tried. The latter valve can be used in this circuit

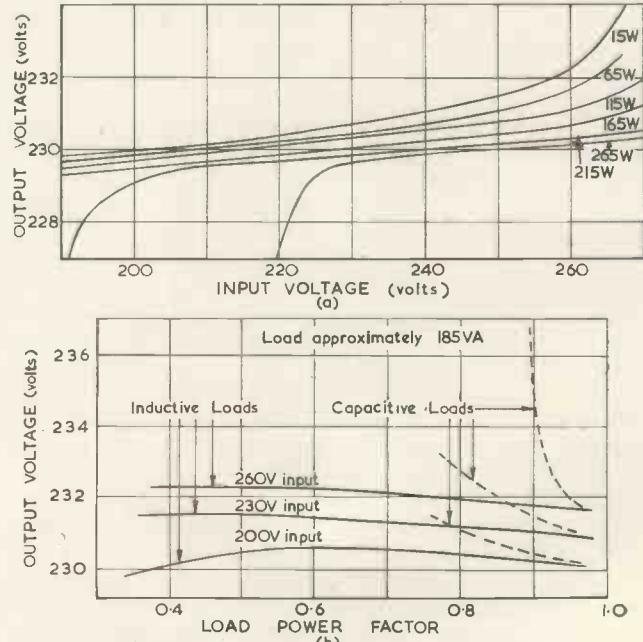


Fig. 3(a). Performance of a.c. stabilizer with resistive loads  
(b). Performance of a.c. stabilizer with reactive loads

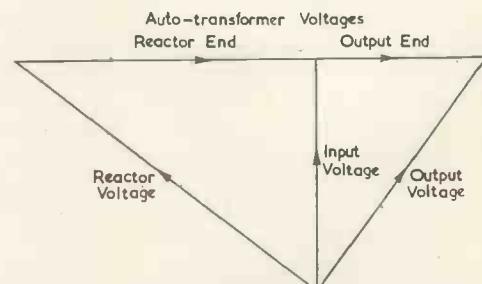


Fig. 4. Typical voltage vector diagram

without other modifications; as it is capable of passing a larger anode current for a given h.t. voltage than the EF55 the maximum output power is greater, but the performance is worse as is shown in Fig. 7.

The feedback circuit, comprising capacitors  $C_2$ ,  $C_3$  and resistor  $R_{10}$ , is used to remove excessive voltages on the electrodes of the amplifier valve. Without the feedback circuit, there are voltage peaks of over 1kV on the anode of this valve, which arise from the 100c/s currents induced in the control winding of the reactor and also a similar component in the emission current of the diode. These voltages, on anode and grid, are considerably reduced by the feedback circuit.

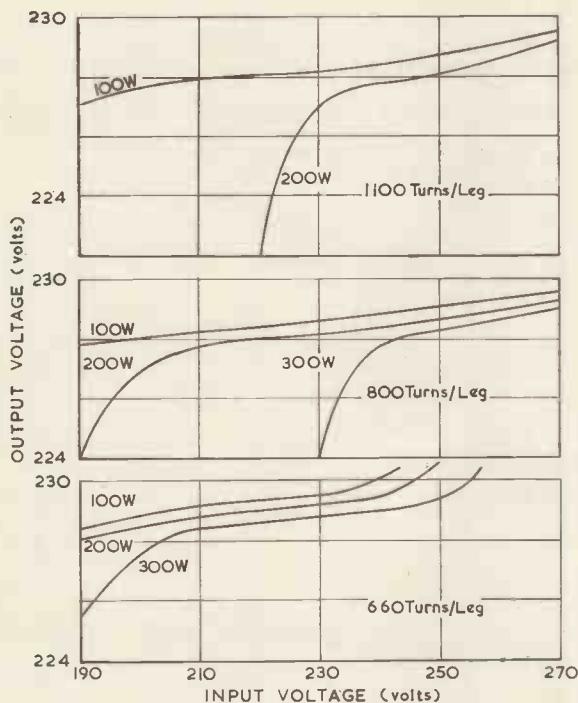


Fig. 5. A.C. stabilizer effect of varying the reactor turns  
Auto-transformer ratio 2.05:1.

Capacitor  $C_3$  is used to block the d.c. path between anode and grid, and it can be varied considerably. Increasing the capacitance improves the response time at the expense of performance without having much effect on the a.c. voltages. Increasing  $C_2$  reduces the a.c. voltages at the expense of response time; if it is too large there may be severe hunting. Should  $R_{10}$  be made too small, dangerous instability may arise, while if it is made too large the unit oscillates in a stable manner. If  $R_{10}$  is reduced to  $\frac{1}{2}\text{M}\Omega$ , there is little risk of instability, but there will be considerable hunting, while if it is much more than  $1\text{M}\Omega$  oscillations may be possible.

It is obviously necessary to have a stable h.t. supply for the diode bridge, since a variation in this voltage would cause the output voltage to vary. It has been found that the output of the unit is sufficiently stable to provide this, providing that steps are taken to prevent transient changes in the output voltage (due to sudden changes in load and/or input voltage) from affecting the bridge voltage. A resistor-capacitor filter with a long time-constant has been found to be sufficient. The time-constant of  $R_{12}$  and  $C_4$  is about  $\frac{1}{2}\text{sec}$ , neglecting the effect of  $C_5$  and the supply impedances.

The variations of voltages and currents in the corrector

circuit with changes of input voltage and load are shown in Fig. 8. As explained previously, voltages on the transformer and reactor will be less with inductive loads and greater with capacitive loads, while the reactor currents increase with inductive loads and decrease with capacitive

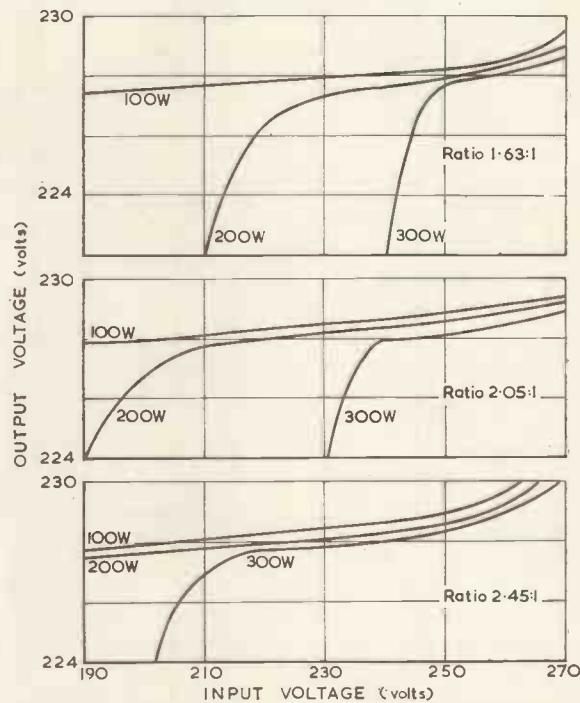


Fig. 6. A.C. stabilizer effect of varying the autotransformer ratio  
Reactor turns 880 per leg.

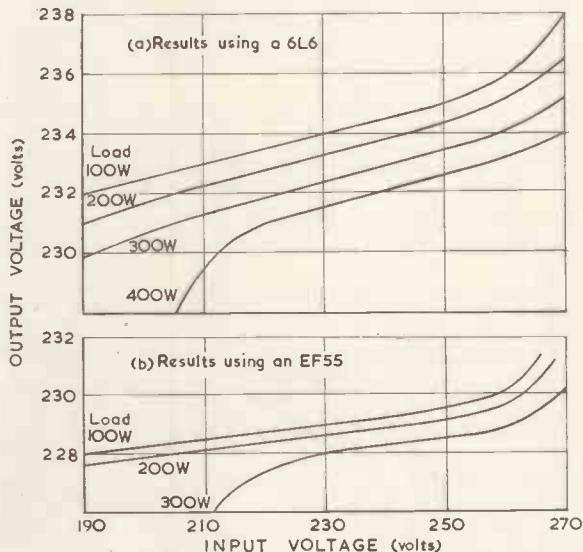


Fig. 7. Stabilizer performance. Comparative performances using different amplifier valves  
Auto-transformer ratio 2.1:1. Reactor turns 770 per leg.

loads, which leads to the limitations of the unit when supplying reactive loads.

In general, with loads having a power factor worse than 0.9 leading, stabilization is not possible, especially with light loads and at high input voltages.

The full output capacity is not, however, always available with lagging loads, although stabilization at the normal

voltage is maintained with power factors down to as low as 0·3, when the maximum available output is 175VA. The performance at low input voltages may fall off somewhat.

The output voltage can be varied by :

- (1) Altering the input tapping on transformer  $T_2$ .
- (2) Altering the filament circuit resistance.
- (3) Altering the diode anode load.

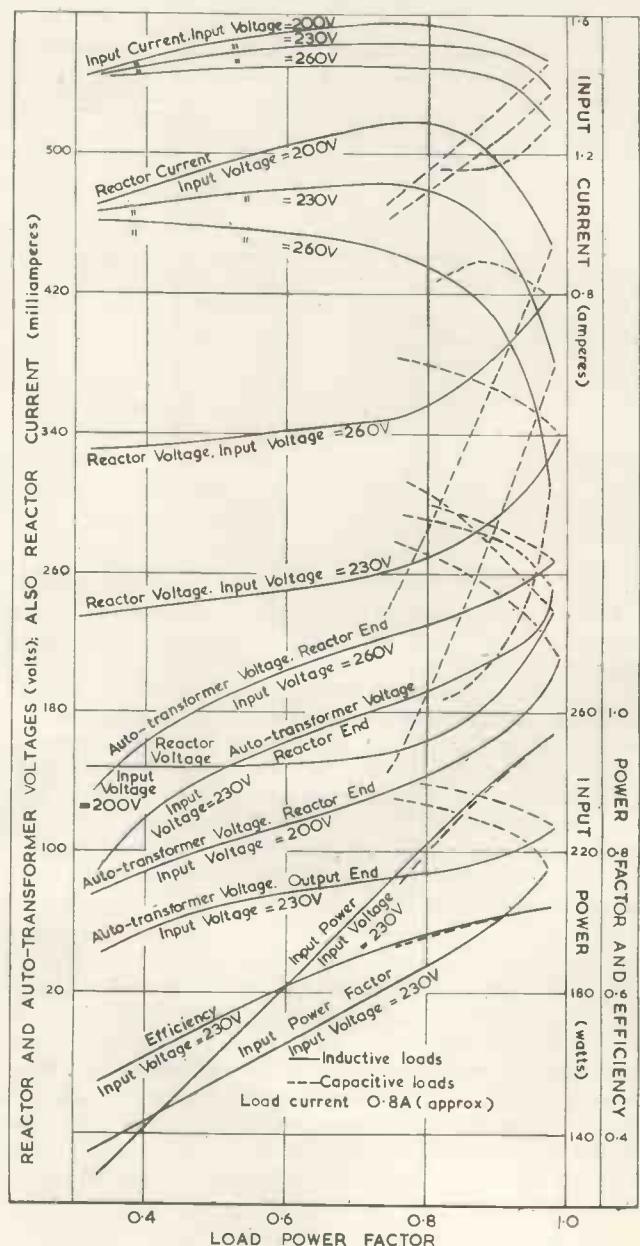
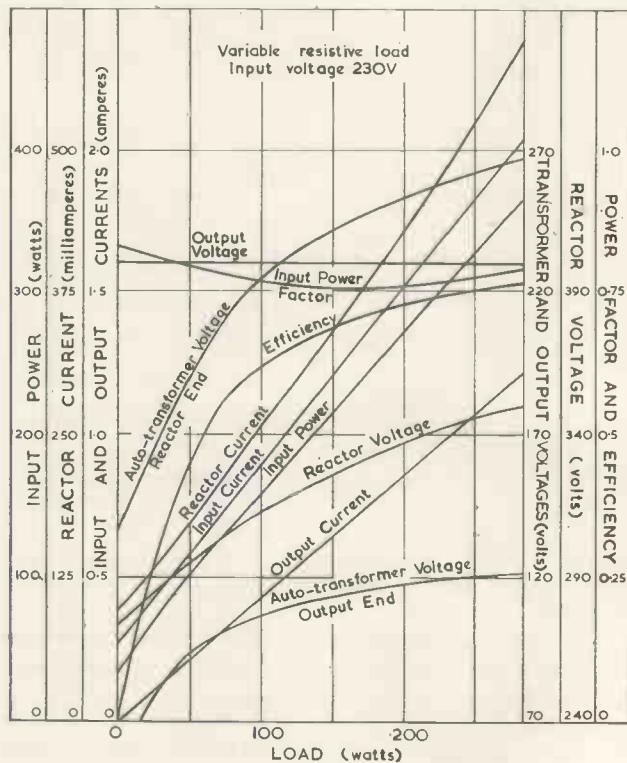
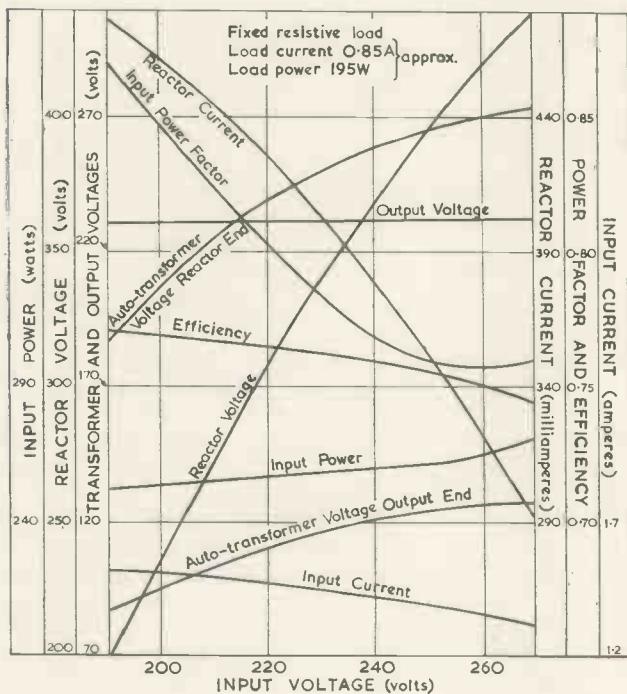
The first is recommended where fairly large variations

are required, while the second and third are recommended to overcome variation of the diode characteristics between valves or through ageing. Resistor  $R_s$  has sections of  $0\cdot05\Omega$ ,  $0\cdot1\Omega$ ,  $0\cdot2\Omega$ ,  $0\cdot3\Omega$ ,  $0\cdot4\Omega$  and  $1\cdot0\Omega$  which can be shorted out as necessary. All connexions on this resistor should be soldered, as it has been found that contact resistance in the ordinary types of variable resistors is not a constant quantity and may give considerable trouble. Variation of the filament circuit resistance by  $0\cdot05\Omega$  changes the output voltage by about 1·5V, while a 10 per cent change in the anode load causes an output voltage change of about 2V.

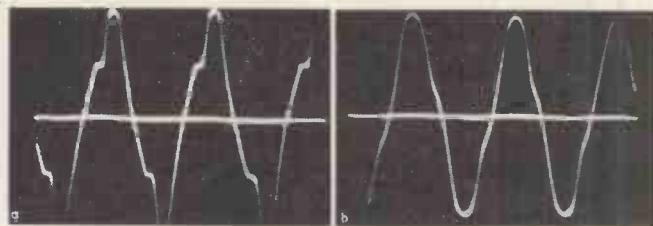
It is advisable to fit a switch on the output of the unit, as otherwise the warming-up period of the unit (normally

Fig. 8. A.C. stabilizer

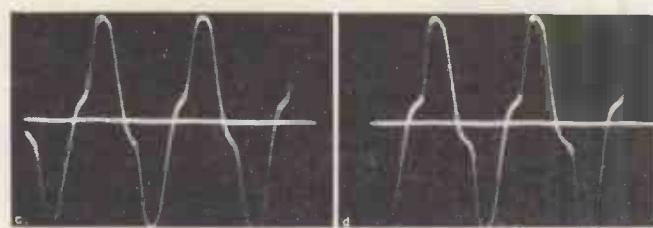
- (a) (top left). Variation of internal voltages, currents, etc., with input voltage
- (b) (bottom left). Variation of internal voltages, currents, etc., with load
- (c) (below). Variation of internal voltages, currents, etc., with load power factor at three input voltages



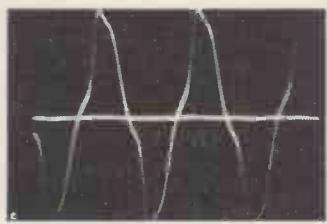
about  $\frac{1}{2}$  min), is considerably prolonged. In addition, excessive voltages appear across the auto-transformer and the reactor when the unit is started up with an appreciable load across its output.



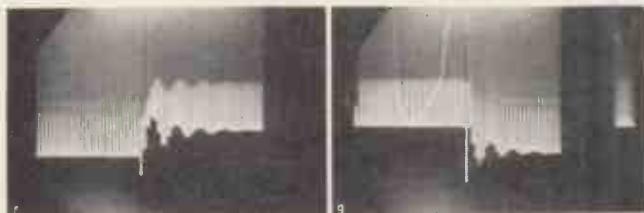
(a) Input voltage 230V. Lagging load about 125W with a power factor approximately 0.6.  
(b) Input voltage 190V. Resistive load.



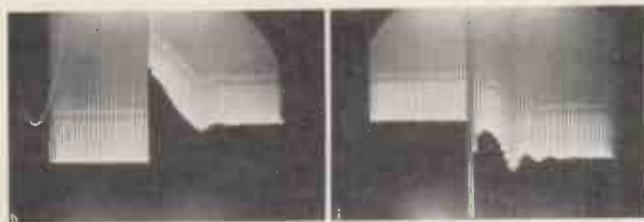
(c) Input voltage 230V. Resistive load.  
(d) Input voltage 270V. Resistive load.



(e) Input voltage 230V. Leading load about 200W with a power factor of approximately 0.9.



(f) Input voltage 230V. Effect of switching on a resistive load of about 0.8A.  
(g) Input voltage 230V. Resistive load of about 0.8A switched off.



(h) Resistive load of about 0.8A. Effect of a sudden change of input voltage from 260V to 200V.  
(i) Resistive load of about 0.8A. Effect of a sudden change of input voltage from 200V to 260V.

Note: In oscilloscopes (f)–(i) only the peaks of the voltage waves are shown.

Fig. 9. Output-voltage waveforms

The output of this circuit is unaffected by changes in supply frequency. Unless the unit is operated near the limits of its stabilizing range, the output voltage will not change by more than a fraction of a volt over the frequency range 40 to 60c/s.

As is usual with circuits using saturable magnetic elements, there is some distortion of the output waveform. Some typical waveforms are shown in Fig. 9, from which it will be seen that the distortion is not usually very large, and should not normally be objectionable.

The time of response of the unit is normally about  $\frac{1}{4}$  sec for sudden changes in either input voltage or load. A slight overshoot is usual with the component values shown.

It is advisable to protect the unit from overloads, which cause excessive voltages on both auto-transformer and reactor should the load be such that the output voltage falls below normal. A fuse of suitable value in the input lead should be adequate; it will also operate in the event of a fault in the unit itself, although an additional fuse in the input line to transformer  $T_2$  might be advantageous.

The efficiency of the unit depends on the load and input voltage. The no-load input is about 45W, and the losses increase with increasing input voltage. With a 200W load, the efficiency is about 70 per cent.

Smoothing of the h.t. supply for the control winding of the reactor has been found to be unnecessary, since the control winding has a large inductance and takes a current which fluctuates considerably—at twice mains frequency. Smoothing by series inductance has shown no beneficial effect, while shunt capacitance may form a resonant circuit with the control winding and lead to dangerous oscillations of the output voltage. Smoothing of the supply to the screen grid of the amplifier valve is, however, highly desirable, as it improves the general performance considerably. The unit is capable of supplying rectifier-fed loads, including those fed by half-wave rectifiers. As might be expected, the waveform distortion is greater than when supplying ordinary resistive loads, especially if the load is smoothed or fed by a half-wave rectifier. Apart from a small reduction in the maximum output available—more noticeable with half-wave and smoothed loads—the performance is not inferior to that obtained with resistive loads.

#### Acknowledgments

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#### REFERENCES

- PATCHETT, G. N. Precision A.C. Voltage Stabilizers. *Electronic Engng.* 22, 371, 424, 470, 499 (1950).
- RICHARDS, J. C. S. A Stabilized A.C. Supply for Lamps and Valve Heaters. *J. Sci. Instrum.* 28, 333 (1951).
- THOMAS, P. A. V. An Alternating Current Stabilizer for Supplying Valve Heaters. *Electronic Engng.* 25, 522 (1953).
- LONG, J. V. P. A Simple Electro-Mechanical Voltage Stabilizer. *Electronic Engng.* 24, 26 (1952).
- COLLINGE, B., MARSHAM, T. N. An A.C. Stabilizer. *J. Sci. Instrum.* 28, 374 (1951).
- HELTHERLINE, L. Diode Controlled Voltage Regulators. *Electronics* 20, 96 (June, 1947).
- British Patent 654 116 (1951).
- ATTREE, V. H. A Differential-Voltmeter Using a Temperature-Limited Diode. *J. Sci. Instrum.* 29, 226 (1952).  
ATTREE, V. H. Measurements of Saturated Diode Stability. *Brit. J. Appl. Phys.* 4, 114 (1953).
- BENSON, F. A., SEAMAN, M. S. Characteristics of the Temperature-Limited Diode Type 29C1. *Electronic Engng.* 25, 462 (1953).  
BENSON, F. A., SEAMAN, M. S. Saturated Diodes—The Characteristics of Some Commercially Available Valves. *Electronic Engng.* 27, 360 (1955).
- JOHNSON, R. P. Construction of Filament Surfaces. *Phys. Rev.* 54, 459 (1938).
- HALL, D. L. Characteristics of Mazda Diode 29C1 (Letter). *Electronic Engng.* 26, 133 (1954).

# BOOK REVIEWS

## Servomechanisms and Regulating System Design

By H. Chestnut and R.W. Mayer. 384 pp. 60 figs. Demy 8vo. Volume II. John Wiley & Sons Inc. New York. Chapman & Hall Ltd. London. 1955. Price 68s.

Although based on Volume I of the same title, which enjoys a well merited popularity in England, the present volume is complete in itself and immediately suitable for a servomechanism engineer already equipped with a sound background of linear feedback system theory and associated mathematics. The authors are now concerned with the application of the basic concepts to actual design problems.

The vital importance of performance measuring techniques is stressed in the first chapter, which deals with open and closed loop testing, in theory and practice. Some useful examples of laboratory test apparatus are described, together with methods of obtaining transfer functions of given servo elements from test results. The production of frequency response data from transient response measurements is also covered.

The important problem, in design, of the effects of noise in the input to a control system is given due weight and a good foundation is laid. It includes an examination of errors arising from input noise and a consideration of "equivalent sinusoidal inputs" which enable the well known attenuation frequency concepts to be applied. The classical statistical approach to noise problems is introduced from a designer's angle, and references for further study are given.

The chapter on the power requirements in a servomechanism contains useful numerical tables, and worked examples based not only on the load characteristics, but on other factors which affect power demands. The important effects of gear ratio on overall performance are brought out clearly.

Since the synthesis of a system usually involves the use of 'shaping' networks to allow given requirements to be met, by producing particular attenuation and phase characteristics, a chapter is devoted to this subject. It contains charts and tables of *RC* networks with a wide variety of transfer functions and should be most useful to designers. Attenuation-frequency ratios are in decibels per decade which may cause initial difficulty to readers already accustomed to decibels per octave, but until a universal unit has been recognized these differences will have to be tolerated, since neither convention is basically superior to the other.

The section on electronic d.c. servo amplifiers is brief, but concise, and concentrates on drift reducing circuits, d.c. level changers, demodulators etc. which

are used in servo design. There is a chapter on all a.c. servos, covering a.c. stabilizing networks and effects of changes in carrier frequency. These details are often neglected in other books.

The final third of the book deals with non-linear servomechanisms and includes the theory of linearization of non-linear elements, their describing functions and the stability of non-linear systems. Numerical examples are given of the effects of saturation, backlash, and load resilience, and the final chapter deals with certain uses of non-linear elements to improve overall performance.

The whole book presents a clear and logical approach to the more advanced aspects of servo design and although electronic systems are taken as examples nevertheless designers of electro-hydraulic and electropneumatic systems will find much to help them. The accompanying diagrams and graphs are excellent and there is an extensive bibliography of relevant books and original papers.

The wide field covered, which is often found scattered throughout several books, helps to make the rather high price, common however to all American books, less of an obstacle to the serious student, who should find it a welcome addition to his servo library.

E. B. PEARSON

## Nuclear Radiation Physics

By R.E. Lapp and H.L. Andrews. 532 pp. 40 figs. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1955. Price 50s.

This book, a descriptive survey of nuclear physics and its application, is to some extent intended for self study and numerous problems are included to this end. The level of mathematics required stops short of calculus for most chapters but a reader should have some familiarity with elementary differentiation and integration to follow the arguments fully.

The treatment, which covers a very wide field, is to some degree governed by historical rather than logical considerations and this has tended to obscure the development in some sections. Against this must be set the undoubted fact that it is desirable, in an introductory text such as this, to give the reader some glimpse of the individual contributions of such men as Rutherford.

A brief first chapter defines elementary physical concepts, such as wavelength and energy and is followed by sections on atomic and nuclear structure. The former deals with the elementary quantum theory of orbital electrons, while the latter describes the neutron-

proton structure of the nucleus and explains the occurrence of isotopes. The energetics of the nucleus are explained very clearly.

A chapter on natural radioactivity precedes those dealing with X and  $\gamma$  rays and  $\alpha$  and  $\beta$  particles. Subjects such as the systematics of  $\alpha$  emitters and the Fermi theory of  $\beta$  decay are well described, although the latter section will tax the comprehension of readers with an elementary mathematical equipment. Treatment of the behaviour of charged particles in separate chapters ( $\alpha$  &  $\beta$  particles in the early part of the book, and protons, mesons, etc., later) obscures the fact that all fast charged particles lose energy to matter by essentially the same mechanisms. In connexion with this subject it is surprising to find in a book revised as recently as 1954 that no reference is made to Cerenkov radiation or to the reduction of stopping power in condensed media. A similar deficiency is found in the next two chapters, on ionization chambers and pulse counters, where only one and a half pages are given to scintillation counters, while an undue amount of space is given to quartz fibre instruments.

Particle accelerators and the nuclear reactions produced by their aid are next discussed. Particle energies produced artificially are beginning to overlap the lower end of the range available in cosmic radiation, and the authors give a most interesting survey of recent work in the field of high energy reactions. Some reference is also made to thermo-nuclear processes.

Owing to their superior ability to penetrate the nucleus, neutrons are generally much more effective in producing reactions than are charged particles. A chapter on neutron physics thus includes much data on the large number of radioactive isotopes produced by neutron capture and provides a clear picture of the status of the transuranium elements. The high neutron density necessary for producing artificial elements in large quantities is, of course, obtained by the use of nuclear fission chain reactors. Nuclear fission and nuclear power are dealt with in two long chapters which present a fairly full picture of the theory of the fission process and the properties of fission products and also of research reactors. Although some data on U.S. power resources is given, no really up-to-date information on power generators is provided and this is a disappointing omission in a book of such general interest.

Three final chapters are on a rather more practical level than those previous and deal with techniques for radiation measurement, the use of radio-active tracers and health physics.

The wealth of basic information distributed throughout the book is backed up by an appendix of useful quantitative data, and adequate references to current literature are given in all chapters.

When one considers the range of phenomena covered in the text, it is not perhaps surprising that a reviewer can

point to omissions or disagree with the emphasis placed on some subjects at the expense of others. One disappointing aspect of the treatment, for instance, is the undue neglect of work outside America, since 1946. This having been said, it is still clear that the authors have achieved their aim of producing an introductory work which has some value as a reference source. It should prove particularly useful to individuals who are newly affected by the growing field of nuclear energy.

J. SHARPE

### An Introduction to Colour Television

By G. G. Gouriet. 72 pp. 28 figs. Demy 8vo. Norman Price (Publishers) Ltd. 1955. Price 8s. 6d.

THE reader must bear in mind that this monograph is a reprint of two lectures, and does not claim to be a comprehensive text-book on the subject of colour television. The footnote to the title and the author's preface both emphasize that it deals mainly with the principles and development of the American NTSC system which, of course, is under test by the BBC at the moment.

The first part of the book deals with Colour Mixture Theory and the second part with the NTSC system. The Colour Mixture section is an abbreviated version (with full acknowledgments) of Professor W. D. Wright's two books "Measurement of Colour" and "Normal and Defective Colour Vision". This condensed version is lucid and adequate but, of course, repeats many of the tortuous conceptions associated with the original C.I.E. colorimetric system. For example, the mixture of colours shown in the colour triangle of Fig. 3 shows the straight line joining two colours  $C_4$  and  $C_5$  divided in the ratio  $d_1:d_2$  to give a mixture colour  $C_6$ . This is only true if the intensities referred to in the text, page 25, are measured in the same units as the reference axes. If the latter are in "T units" to ensure a white point at the centre of the triangle, then the ratio  $d_1:d_2$  is not the ratio of the luminosities of  $C_4$  and  $C_5$ —a point which the practical user of the C.I.E. colour triangle soon meets rather forcibly.

In the second part, the author maintains his high standard of lucidity in dealing with the NTSC system. The field sequential system is disposed of in a page, as are "Other Possibilities" at the end. With regard to the latter, the promise in the Preface in this section "These (proposed methods of instrumentation) and many other aspects of the (NTSC) system are dealt with comprehensively ..." is, unfortunately not realized.

The section 2.6 "The Failure of Constant Luminance" is mainly of academic interest. In practice this turns out to be the least of the criticisms levelled at the NTSC system. With regard to the last

paragraph in section 2.8, page 66, it is generally agreed "that the NTSC system must surely be regarded as an outstanding achievement" but not so generally that it is "a first-class example of how a communication channel may be used efficiently".

The illustrations are excellent, but it is a great pity that the printing industry continues to misuse subtractive primary colours. This comment applies particularly to Fig. 7 the implication of which is confused by the failure to reproduce a good healthy red and cyan in the appropriate parts of the diagrams.

This book will appeal to all television engineers who are beginning to take an interest in colour television, and the references contain some good suggestions for further study.

L. C. JESTY

### Éléments de Mécanique Quantique

By P. Pluviniage. 547 pp. 79 figs. Demy 8vo. Masson et Cie. Paris. 1955. Price 4 600 Fr.

THIS book gives a well balanced survey of elementary quantum mechanics, is an informative and valuable addition to the literature and offers a stimulating approach to the subject. Professor Pluviniage recognizes the difficulty the student encounters when passing over the French traditional instruction methods to what he terms the methods of "stirring physics". The author places himself at the level of a student who has already followed courses in general physics and mechanics, and attempts to level out the radical change of spirit and outlook involved. It is courageous to have undertaken to outline these fundamental ideas in so short a compass, the main problem confronting the author has clearly been that of selection. He has been most successful in solving it, and students and teachers will be grateful to the author for devoting his special gifts to this task.

Part 1 deals with particle movement in an axis, detailing the classical mechanics of the particle, Schrödinger's representation and equation, the harmonic oscillator, discontinuous representations and the Heisenberg theory. Part 2 considers the particle in space, making a special study of curvilinear co-ordinates and uncertainty in classical mechanics, orbital momentum and spherical harmonics, the vibrational rotational spectrum of diatomic molecules, the normal Zeeman effect and the Paschen effect. There follows a third part which deals with notions about particle systems, giving a brief but adequate account of the general principles of quantum mechanics in relation to particle systems, diatomic molecules and the semi-quantum theory of electromagnetic radiation. The text has been written in a most attractive style, and one is left with the impression that the student using this book would realize that his steps were being guided by a master of the subject.

E. LEWIS

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# The Physical Society's 40th Exhibition

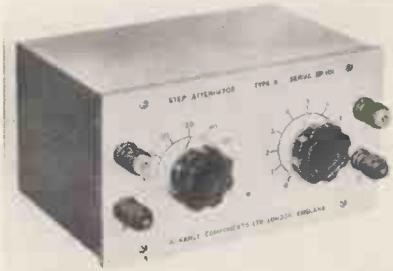
A description, compiled from information supplied by the manufacturers, of selected exhibits at the Physical Society's Exhibition held in London from 14th to 17th May.

## Advance Components

### L.F. Attenuator

(Illustrated below)

**T**YPE A64 is a variable attenuator with a constant input and output impedance of  $600\Omega \pm 1$  per cent. Designed for use with audio frequency apparatus and also having applications in telecommuni-



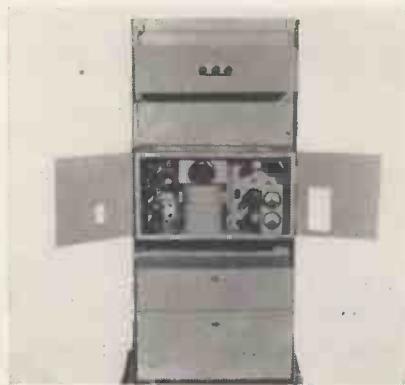
cations, it can be used up to at least a frequency of 1Mc/s, where the errors do not exceed 2.5dB. The A64 covers a range of 70dB and consists of a section with six steps of 10dB followed by a fine control of ten 1dB steps. The single steps of 1dB and the first step of 10dB are each independent from all other positions. The remaining 10dB steps are part of a ladder network compensated to maintain constant impedance. The attenuation is reversible and a zero loss position is provided.

**Advance Components Ltd,**  
Roebuck Road,  
Hainault,  
Essex.

## Airmec Standard Frequency Equipment

(Illustrated below)

**T**HE total equipment consists of a number of units which may be supplied individually if required, and can be most conveniently considered as four standard rack assemblies. Rack A normally contains the crystal oscillator and associated power unit and distribution panels. Rack B contains frequency divider units, a clock unit and



comparison equipment when required. Racks C and D contain protection equipment consisting of auxiliary power supplies and changeover units.

The oscillator employs a 100kc/s Essen ring type crystal in an oven, the temperature of which is controlled to within very close limits. In general the yearly frequency drift should not be more than 1 part in  $10^7$  and yearly stabilities of up to 1 part in  $10^8$  are not uncommon.

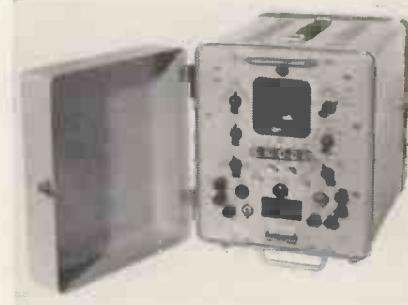
**Airmec Ltd,**  
High Wycombe,  
Buckinghamshire.

## British Telecommunications Research Telegraph Distortion Measuring Sets

(Illustrated below)

**S**ETS TDMS.5a and TDMS.6a provide comprehensive testing facilities for line and radio teleprinter circuits and equipment. The two units, which are portable, may be used independently or in combination. Alternative front panels are available to allow for two units to be mounted side-by-side on an international 19in rack.

TDMS.5a is basically a transmitting unit which supplies test signals at any speed from 20 to 80 bauds, the signals being either perfectly timed or given any degree of start element distortion up to



$\pm 50$  per cent and of 7,  $7\frac{1}{2}$  or 8 unit code. Six types of signal can be generated: one-to-one reversals, two-to-two reversals, steady mark, steady space, repetition of a signal character as selected by five keys on the front panel and a 100-character test message from an electro-mechanical coder contained within the instrument.

The outgoing signals are presented on a cathode-ray tube with a circular time-base and this display may also be switched to a telegraph line to provide for measurement of distortion on synchronous teleprinter signals or measurement of relay bias.

TDMS.6a is the receiving unit used to measure the distortion of either synchronous or start-stop signals. It can be used on working circuits without interrupting the traffic. For synchronous signals the time-base is circular, as in unit

TDMS.5a and the same facilities are given. For start-stop signals a start-stop spiral time-base is used, there being one revolution of the spiral for each element of the character, from 4 to 14 elements. The time-base is triggered by the start element of an incoming character and each succeeding instant of change-over causes a bright dot to appear on the spiral trace allowing for measurement of each element length. The speed of transmission is found by adjusting the time-base speed to obtain a stationary train of dots and reading the speed directly from the controls. Speeds of from 20 to 80 bauds can be measured with an accuracy to  $\pm 0.5$  baud.

**British Telecommunications Research Ltd,**  
Taplow Court,  
Taplow,  
Nr. Maidenhead,  
Berkshire.

## B.T.H.

**X-Band Precision Waveguide Attenuator**  
**T**HIS attenuator has a calculable law of attenuation over a wide frequency band to an accuracy of at least  $\pm 0.1$  dB with a voltage standing wave ratio better than 0.95. The insertion loss is less than 0.25dB and the phase shift is small.

More or less of the incident power is absorbed by rotating an absorbing fin about the longitudinal axis of a circular waveguide transmitting a linearly polarized wave.

## Frequency Comparator

This equipment is for frequency comparison of high stability quartz crystal oscillators, which never differ by more than 9c/s, for transient fluctuations and long-term drift in frequency. One oscillator output is continuously phase-shifted by a servo-motor and the two frequencies are demodulated to d.c. which is displayed showing the transient fluctuations. The angular velocity of the servo-motor output shaft is converted to electrical pulses which are counted every 15 minutes, hence the drift from one 15-minute period to the next can be ascertained.

**The British Thomson-Houston Co. Ltd,**  
Rugby.

## A. E. Cawkell Servoscope

**T**HIS is a comprehensive instrument designed for waveform examination in the frequency range 0.1c/s to 20kc/s.

Two high gain d.c. amplifiers connect to a double gun long persistent cathode-ray tube. Sensitivity is 3.5mV/cm maximum, and 6dB-step attenuators are used for gain control. The amplifiers have very low drift and hum levels, and the inputs are balanced differential or single sided. Gain calibration is achieved by

incorporating a rectangular waveform generator producing a voltage output which, connected to the amplifiers, checks their gain and characteristics. The same generator produces marker pips at a number of different time intervals for time calibration.

The time-base provides linear sweeps in the range 0·01c/s to 10kc/s, calibrated with coarse and fine controls.

Beam blanking facilities are fitted and all power supplies are stabilized.

#### Multiple Pulse Generator

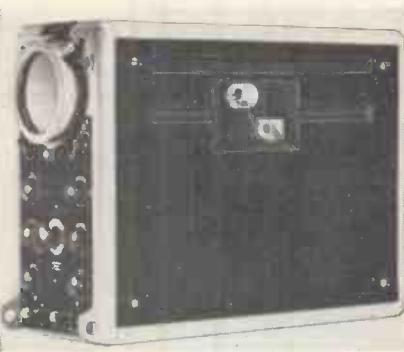
A Dekatron trigger generator continuously generates a "ring of ten" triggers. The ten Dekatron cathodes are connected to an exchange board, where they may be connected in any combination to the three pulse generators and provide a trigger pulse for operating an oscilloscope. The oscilloscope may be triggered from any of the trigger pulses feeding the three generators via a fixed delay of 0·2μsec so that the leading edge of the pulses may be examined. The generators produce pulses of height variable from 0 to 100V and 1 to 200μsec long. The outputs are combined in a mixer circuit which drives a high power cathode-follower stage. External capacitance of up to 0·005μF may be connected and the output pulse waveform will rise to 100V positive or negative in 1μsec. The output impedance is 25Ω.

With suitable adjustments on the exchange board and the height and length controls, patterns of negative and positive pulses of variable height in recurring combinations of up to 10 may be obtained from a single output terminal.

**A. E. Cawkell,**  
6-7, Victory Arcade,  
Southall,  
Middlesex.

**Cossor Instruments**  
**Milli-Microsecond Pulse Generator and**  
**Display Unit**  
(Illustrated above right)

THE type 1060 has been designed to test passive and active networks by pulse technique and comprises an adjustable pulse source, a display oscillograph and power units. A free-running blocking oscillator and shaping circuits generate pulses of approximately 100V peak amplitude and of substantially "sin" shape, adjustable in width to 5, 20 or 50μsec. The oscillator provides also the sawtooth voltage for the oscilloscope time-base so that the transmitter pulse is always presented at the start of the cathode-ray traverse and automatically locked. If a passive network—such as a mismatched cable (which may vary in length from about 18in up to approximately 200 yards)—is connected to the pulse source, reflections of the transmitter pulse will appear on the transverse as echoes. From the characteristics of the echoes, a large amount of information about such a cable is learned, a discontinuity—with its location and magnitude—being readily ex-



posed. The cable impedance and its electrical length are measured and its losses and phase response determined.

**Cossor Instruments Ltd,**  
Highbury Grove,  
London, N.5.

#### Decca Digital Computer

THE Decca C1 computer is a high speed serial digit computer intended primarily for computation on scientific problems. Input information is fed to the machine by punched tape prepared on a teleprinter which simultaneously prepares a copy in plain language. Programme input is in the form of words, each of 40 binary digits. Each word specifies two operations and includes the address of the next instruction. For arithmetic operations a decimal notation is adopted and a word then consists of a digit indicating the sign followed by a 9 decimal digit word.

Information within the computer can be stored either on a 63 channel magnetic drum or in magnetic core shift registers when quick access is required. A novel technique is used in the magnetic drum store on which the digit rate is three times that of the computer itself. The access time to the drum is thus one-third of what it would be if the drum rotated at a speed appropriate to a 1:1 digit rate.

Decca C1 employs magnetic core circuits to perform not only storage and control functions but also to perform arithmetical operations. The cores used in the C1 are ring-shaped and are only 2mm in diameter. Consequently, the power required to operate them is very low.

**Decca Radar Ltd,**  
1-3, Brixton Road,  
London, S.W.9.

#### Dynatron Radio Dekatron Scaling Unit

THE type 1287A is a portable five decade scaler with built-on Geiger-Muller tube and sample chamber. The detector, comprising a halogen quenched g.m. tube with an annular ring type of sample holder is followed by a five decade scaling circuit. The circuit uses five Dekatron tubes, enabling a total of 99 999 counts to be registered. The timing circuit incorporates a stop-watch and counting can be either "count-controlled" or "manual (time) controlled".

Power is derived from a 4·8V 7Ah "Nife" battery and self-rectifying vibrator, 10 hours' operation from each discharge cycle of the battery being obtained.

The 1287A incorporates a pulse generator providing fast or slow "self test" facilities at the operation of a spring loaded switch. Built-in and controlled from the same switch is a small meter indicating battery voltage.

**Dynatron Radio Ltd,**  
The Firs,  
Castle Hill,  
Maidenhead,  
Berkshire.

#### Edison Swan V.L.F. Function Generator

THIS instrument has a continuously variable frequency of 0·01 to 1 000c/s. The output is available simultaneously as a sine, square and triangular waveform, with a peak-to-peak voltage of up to 100V from cathode-follower stages. Alternatively, any one of the waveforms is available push-pull into 600Ω with a peak-to-peak value of up to 50V. An attenuator is fitted.

A feature of this instrument is that there is no transient amplitude change when the frequency is altered.

Lower frequencies can be obtained by the addition of external components.

**The Edison Swan Electric Co. Ltd,**  
155, Charing Cross Road,  
London, W.C.2.

#### Ekco Electronics

##### Auto-Standardization Thickness Gauge

THE Ekco auto-standardization thickness gauge automatically and periodically removes the measuring head from the material being tested and makes its own adjustments to correct for source decay, contamination and amplifier drift, etc.

Automatic scanning facilities permit checking the material across its entire width and on the occurrence of a fault in the equipment or breakdown in the manufacturing process, the equipment automatically shuts down and withdraws the measuring head.

#### Reactor Instrumentation

In addition to a variety of standard Ekco instruments which are directly suitable for reactor instrumentation, entirely new instruments are being developed, including Reactor Period Meters, a Power Error Meter and a Fast Neutron Monitor.

**Ekco Electronics Ltd,**  
Southend-on-Sea,  
Essex.

#### Elliott Micro-Microammeter

THIS d.c. amplifier (type ND12) measures very small direct currents, such as those obtained from ionization chambers, has six ranges with full-scale values from  $10^{-11}$  to  $10^{-6}\text{A}$  and is

designed for maximum reliability. It incorporates a sub-miniature electrometer triode, two magnetic amplifier stages and an internal stabilized power supply for ion chamber polarization. To ensure maximum stability the electrometer valve and high-value resistors are housed in a sealed and desiccated unit which is detachable and may be operated remotely from the instrument.

The output is suitable for the direct operation of pen recorders. Provision has been made for a further magnetic amplifier unit to give a warning if the current exceeds a preset level.

**Elliott Brothers (London) Ltd,**  
Century Works,  
Lewisham,  
London, S.E.13.

#### English Electric Magnetrons

THE M525 is for pulse working at frequencies in the S band and with a pulse input of 70A at 37kV will give an r.f. output of approximately 1½MW.

The valve differs from previous valves designed for this rating in that it will work with a rate of rise of voltage pulse as high as 200kV/μsec and requires a much smaller permanent magnet. The magnetic field requirement is 1800G across a 2in gap.

The M541 is designed for pulse working at frequencies in the L band, is mechanically tunable over a 10 per cent frequency band and conforms to the American specification for the 5J26.

#### Transmitting Valves

In the transmitting valve section is the CR1101 u.h.f. tetrode capable of taking full ratings up to a frequency of 900Mc/s. The air-cooled radiator is integral with the heavy walled copper anode and can dissipate 2kW.

At 900Mc/s, a c.w. output of 600W is obtained with a circuit efficiency of 75 per cent. In television service (CCIR) a peak synchronizing power of 1.2kW can be obtained. At 400Mc/s the valve will give at least 1kW useful output.

The most interesting technical feature of this valve is the use of ceramic instead of glass for the vacuum envelope. The ceramic is of the high alumina type and is joined to the metal parts by brazing, having first been metallized.

**The English Electric Valve Co. Ltd,**  
Waterhouse Lane,  
Chelmsford,  
Essex.

#### Ericsson Vibrating Reed Electrometer

(Illustrated above right)

IN nucleonic work it is often necessary to measure low voltages and small currents which originate in high impedance sources such as ionization chambers. This instrument uses the principle of a vibrating capacitor acting as a d.c./a.c. convertor giving a high input impedance which, in the type 126 is better than  $10^{14}\Omega$ . The convertor is followed by a

negative feedback amplifier and a phase discriminating detector to give a high degree of stability, the indicating meter reading varying by less than 100μV in 24 hours.

**Ericsson Telephones Ltd,**  
Beeston,  
Nottingham.



Ferranti

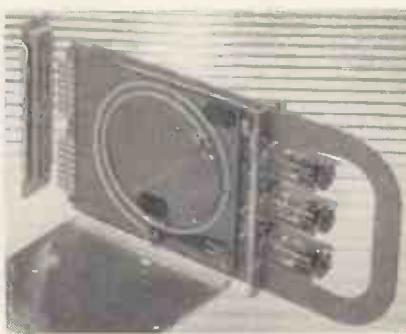
#### Long and Short Nickel Delay Line Stores (Illustrated below)

THESE are used for information storage in digital computers, and two forms are exhibited.

The smaller unit is in general used for the storage of a single "word" for immediate access; the example shown has a capacity of 42 binary digits at a period of 3μsec. The acoustic pulses are transmitted in the longitudinal mode along a coiled line consisting of a bundle of nickel strips.

The other unit is for use as part of a main store in a computer, and consists of a spiralled delay-line constructed from a temperature stable alloy, along which torsional acoustic pulses are propagated, a storage capacity of the order of 1 200 binary digits at a period of 3μsec being obtained.

**Ferranti Ltd,**  
Hollinwood,  
Lancashire.



#### G.E.C. Pulse Code Modulation Pattern Generator

THIS equipment provides an arbitrary ten-digit pattern of binary pulses at a digit rate of 1Mc/s and can be used as a computer storage unit.

Counting-down circuits provide pulses at a repetition frequency of 100kc/s, fed to a magneto-strictive delay line with ten output coils spaced at single digit intervals along its length. Switches on

the front panel of the unit control the output from these coils so determining the pattern produced when the outputs are combined.

**The General Electric Co. Ltd,**  
Wembley,  
Middlesex.

#### Marconi Instruments Carrier Deviation Meter

(Illustrated below)

THE type TF791C direct-reading deviation meter covers a frequency range of 4 to 250Mc/s at input levels of 55mV and above; frequencies up to 500Mc/s can be accommodated at increased input levels. Suitable for both communication and broadcast systems, it has a modulation frequency range of 50c/s to 35kc/s and four deviation measurement ranges of 5, 25, 75 and 125kc/s full scale. A high order of measurement accuracy is ensured by the provision of a crystal controlled deviation standard.

**Marconi Instruments Ltd,**  
St. Albans,  
Hertfordshire.



Mullard

#### Backward Wave Oscillator and Travelling Wave Tubes

IN the backward wave oscillator, as in the travelling wave tube, a beam of electrons is passed through a slow wave structure. Interaction takes place between the beam and a wave travelling along the slow wave structure at about the same speed, causing the beam to deliver energy to the wave.

The backward wave oscillator is so called because the signals generated travel in the opposite direction to the electrons in the beam.

A feature of the b.w.o. is the enormous frequency variation which can be obtained by electronic tuning. The frequency of oscillation is a function of the beam velocity, and hence is determined by the e.h.t. voltage. The Mullard b.w.o. has a tuning range of 11 500Mc/s to 18 000Mc/s, and a frequency sensitivity of about 5Mc/s/V.

Possible application for the b.w.o. are wide band microwave panoramic receivers, wobbulators and other microwave test gear.

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



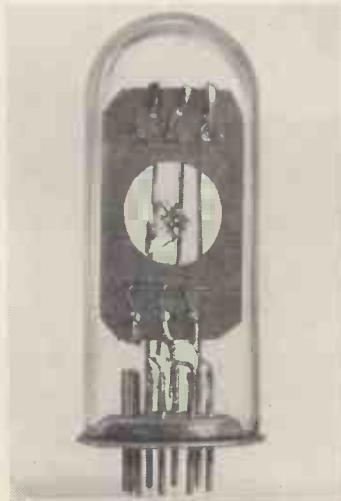
**Muirhead**  
**Phasemeter**  
(Illustrated above)

THE phasemeter type D729 is an instrument for measuring the phase-angle and voltage difference between two sinusoidal voltages derived from the same source, e.g. the gain and phase change of an amplifier, without imposing any appreciable load on the circuit under test. Used with a D699 Frequency Analyser, accurate measurements on distorted waveforms may be made.

**Muirhead & Co. Ltd,**  
Beckenham,  
Kent.

**Nash and Thompson**  
**Shorted Turn Detector**

THE coil or transformer winding to be tested for shorted turns and continuity before insertion of its core is placed on a jig, so coupling it to the tuned circuit of an oscillator. A shorted turn causes damping of the oscillator, shown by the change of reading on a meter incorporated in the instrument. Continuity is tested by deliberately short-circuiting the ends of the winding and is again indicated by a change of reading on the meter. A range of jigs is available to suit a large number of coil sizes. With the correct jig, one shorted turn of 40 s.w.g. can be detected in the outside



layer of a coil of  $\frac{1}{4}$  in radial winding depth.

**Nash and Thompson Ltd,**  
Chessington,  
Surrey.

**Plessey**  
**Single Crystal Memory Cell**  
(Illustrated below left)

THE single crystal memory cell comprises a crystal of barium titanate with suitable electrodes and it gives electric displacement—(field loops with squareness) ratios of the order of 90 per cent. Such crystals may be used as binary stores and have low power consumption and small size.

**The Plessey Co. Ltd,**  
Ilford,  
Essex.

**Salford Electrical Instruments**  
**Linear Function Sub-Miniature**  
**Potentiometer**

(Illustrated below)

THE smallest of a series of linear function sub-miniature potentiometers features a special resistance card and clamping ring construction. No arrangements have been provided for ganging, but the single unit measures only  $\frac{1}{8}$  in



diameter by  $\frac{1}{2}$  in long, not including terminal projections. The extremely small dimensions make this potentiometer suitable where space is limited and where accuracy must be maintained.

**Salford Electrical Instruments Ltd,**  
Silk Street,  
Salford, 3,  
Lancashire.

**W. H. Sanders**  
**Signal Generator**

THE wide band microwave signal generator type XT.312, incorporating a plug-in klystron, gives continuous frequency coverage from 1.5 to 4.5 Mc/s.

The instrument covers approximately a 3:1 frequency range, and contains built-in power monitor, piston attenuators, modulator unit and power supply.

**W. H. Sanders (Electronics) Ltd,**  
Gunnel's Wood Road,  
Stevenage,  
Hertfordshire.

**Servomex**

**Low Frequency Waveform Generator**  
(Illustrated above right)

THIS instrument (type L.F.51) can generate sine waves at frequencies up to 500 c/s and down to one cycle per 2 000 seconds. Eighteen different types of



impulse function are also available, i.e., triangular waves, sawtooth waves and pulses.

**Servomex Controls Ltd,**  
Crowborough Hill,  
Jarvis Brook,  
Sussex.

**Solartron**  
**Mechanical Reference Generator**

TOGETHER with a transfer function analyser set-up, the mechanical reference generator type JX.603 has a wide scope in the fields of mechanical servo systems and automatic controls. By using a synchro attached to the mechanically moving part, the generator creates an electrical reference of input, which is compared with the signal at the output end of the mechanical system, so giving a true measurement of the performance of the servo system.

**The Solartron Electronic Group Ltd,**  
Thames Ditton,  
Surrey.

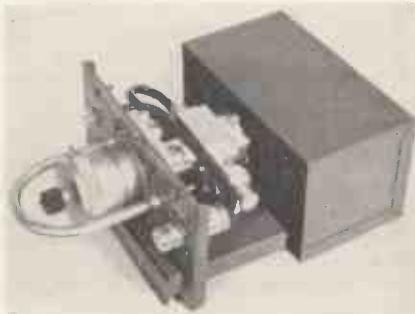
**Wayne-Kerr**  
**Resonator Cavity Wavemeters**

(Illustrated below)

A SERIES of resonator cavities operating in the E0<sub>10</sub>-EH hybrid mode give frequency coverage of 1 000 to 40 000 Mc/s. Each cavity is complete with its connectors and calibration chart. The resonators are free from undesired responses. When necessary, low-pass filters are fitted to attenuate responses above the frequency range of the cavities. These filters give a minimum of 40dB attenuation in the stop-band.

Accuracy of frequency measurement is better than 1 in 3 000, but can be as high as 1 in 10 000 if the coupling is reduced.

**The Wayne-Kerr Laboratories Ltd,**  
Sycamore Grove,  
New Malden,  
Surrey.



## Short News Items

The President of the Institution of Electrical Engineers, Sir George Nelson, recently presented the Certificate of Honorary Membership to Colonel Sir Arthur Angwin, K.B.E., D.S.O., M.C., T.D., a past president of the Institution, in recognition of his outstanding life's work in the field of telecommunication, and of his distinguished services to the Institution. At the same meeting the Faraday Medal was presented to Emeritus Professor G. W. O. Howe, D.Sc., LL.D., M.I.E.E., for his pioneering work in the study and analysis of h.f. oscillations and on the theory of radio propagation, and for his outstanding contributions to engineering education.

An International Congress of Cybernetics will be held in Namur, Belgium, from 26-29 June. This is being organized under the patronage of the Ministry of Education and of U.N.E.S.C.O. and will be under the honorary presidency of the Governor of the province of Namur. The Congress will be presided over by Professor George R. Boulanger of the Polytechnic Faculty of Mons and Brussels University. Further details may be obtained from the Secretariat du Congres International de Cybernetique, 13 rue Basse-Marcelle, Namur, Belgium.

The wind-tunnel of the Aircraft Research Association at Bedford was recently officially opened by H.R.H. the Duke of Edinburgh. In this tunnel is installed a Pye industrial television camera, enabling scientists to observe remotely the characteristics of models undergoing tests in transonic conditions.

A new large transonic testing installation, the 8ft by 6ft transonic wind tunnel, was recently opened at the Royal Aircraft Establishment, Farnborough, by the Minister of Supply, Mr. R. Maudling. The installation is a modified and improved form of the R.A.E.'s well-known "High Speed Wind Tunnel" which began work during the war. Until the autumn of 1954, most of the test work in this country on models of high-speed aircraft at high subsonic speeds was done in it. The total power of the installed machinery has now been increased fivefold.

At the annual general meeting of the British Electrical Development Association, Viscount Chandos, D.S.O., M.C., was re-elected president for the year 1956/57.

An Electronic Exhibitions Joint Association Committee has been formed and is actively considering the whole question of national exhibition policy in the industries concerned. Associations represented are the British Electrical and Allied Manufacturers' Association, British Radio Valve Manufacturers' Association, Radio Communication and Electronic Engineering Association, Radio and Electronic Component Manufacturers' Federation. The RCEEA Secretariat is acting for the Committee from 11 Green Street, London, W.1.

The Society of Instrument Technology Ltd, founded in 1945, now has a membership of nearly 1400. Through the courtesy of the Council of the Scientific Instrument Manufacturers' Association of Great Britain Ltd, an office has been taken at 20 Queen Anne Street, London, W.1, under the secretaryship of Commander A. A. W. Pollard.

The British Council are arranging 25 courses, with places for more than 700 overseas specialists, in 16 centres in the United Kingdom during the current year. A course on Non-destructive Testing of Materials will be held in London from 4-17 November. Further details on all courses, covering arts and humanities, education, medical, science and social science, may be obtained from the British Council, 65 Davies Street, London, W.1, or through British Council offices overseas.

The University of Birmingham Department of Electrical Engineering and Department of Extra-Mural Studies are arranging a summer school on Communication Theory, Modulation and Noise to be held from 1-13 July. It is intended for engineers in the electrical communications and allied fields who have taken their degrees without much theoretical work in communications or feel the need for a "refresher" course. The treatment of the subjects will be basically quantitative, although it is not intended that the mathematics involved should be very advanced. There will be no practical work, but students will be encouraged to spend some spare time in the research laboratories of the Department. The fee for the course will be £8. Residential accommodation can be provided, for a maximum of 50, at Chancellor's Hall, Edgbaston, Birmingham 15. The charge for the 12 nights, including breakfast and dinner, will be £12. Applications for the course should be made to the Director of Extra-Mural Studies, The University, Edmund Street, Birmingham 3.

Marconi's Wireless Telegraph Co Ltd have been awarded the contract for the construction of the Independent Television Authority's second Northern television station. This is to be built at Emley Moor, near Huddersfield. The Marconi Company will be responsible for the entire station, comprising the buildings, transmitters, tower, aerial and feeder systems.

The Middle East's first television station was recently opened in Baghdad. During the early stages Pye Ltd are to provide experienced engineers to advise the Iraq personnel on the running of the station. Since Pye sold the television station to the Iraq Government, it has been removed from the site on which it stood during the Baghdad Fair, and has been re-erected on a permanent location from which it will have an initial range of ten miles, sufficient to cover the whole of Baghdad. Later a high powered transmitter will be installed to increase the range.

"FINAC" is the name which has been given to the first electronic computer to be installed in Italy for the Instituto Nazionale per le Applicazioni del Calcolo. Manufactured by Ferranti Ltd, I.N.A.C. are to use the computer for operating a computing service for Italian industry. Organizations will bring their particular problems to the Institute, and provided they are of a suitable kind the Institute will arrange for them to be "programmed" for the computer. Or in some cases, such as a straightforward solution of simultaneous equations, a programme may exist already.

The United Kingdom Atomic Energy Authority has issued a list of publications which are available to the public. The depository libraries from which they may be borrowed are as follows. The Science Museum Library, South Kensington, London, S.W.7. Central Library, Ratcliff Place, Birmingham 1. The Mitchell Library, North Street, Glasgow, C.3. Central Library, William Brown Street, Liverpool, 3. Central Library, St. Peter's Square, Manchester, 2. Central Library, New Bridge Street, Newcastle-upon-Tyne, 1. Central Library, Surrey Street, Sheffield, 1.

**Erratum.** In Table 1 of the article "Tolerance Limits in Matching" (April issue, page 163), all the values of power transfer are negative and not " $\pm$ " as shown.

# LETTERS TO THE EDITOR

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

## A D.C. Coupled Circuit Using Voltage Stabilizing Valve

DEAR SIR.—The type of circuit described by Mr. Court (December, 1955 issue) has been employed for exactly the same purpose in an a.f.c. system developed by my company. We find that the subminiature CV2213(NT2) neons perform this task adequately, require no trigger resistors, and drain only 0.6mA from the negative line. It is necessary, however, to by-pass the whole chain with a decoupling capacitor because neon valves generate noise and are subject to modulation by stray r.f. fields.

In my opinion the chief merit of the method is not that it avoids attenuating the control signal but that the neons take over one function of the a.f.c. system, namely stabilizing the klystron frequency against supply variations. If all the klystron and a.f.c. supplies are left unstabilized and a suitable proportion of the coupling chain is stabilized the net variations of the klystron electrodes can be designed to have just zero effect on its frequency, leaving the entire error-handling capacity of the a.f.c. system to cope with disturbances from other sources.

If this feature is not important, it is worth noting that there may be an even simpler solution to the problem, which avoids attenuating the control signal at all. If, in Fig. 2 of Mr. Court's article the 600V line is increased to 750V but made a floating supply with its positive end connected to the pentode anode, then anode variations can be picked off at any potential down to -600V. The drawback is that the 600V line may be employed elsewhere where these variations are objectional, but we find that the normal supply for t.r. cell "keep alive" electrodes (actually -1kV) can be floated in this way for the dual purpose.

Yours faithfully,

R. J. D. REEVES

E. K. Cole Ltd.  
Malmesbury,  
Wilts.

## The author replies :—

DEAR SIR.—I was interested to read Mr. Reeve's comments and to note the techniques used by E. K. Cole Ltd.

Since our equipment was experimental, neons readily available were used, although it is apparent that in the case of airborne equipment the use of subminiature types could be advantageous.

The argument in the second paragraph is not clear and possibly Mr. Reeves may care to enlarge upon it. In fact, our supplies being used for other parts of the circuit, as is normally the case, were stabilized and the balancing out of voltage variations on the klystron reflector

did not arise. The circuit was used, as stated, in order to avoid attenuation of the control signal.

In our application the provision of a floating 750V power supply does not appeal as being easier than the method adopted. Since the klystron cathode is supplied with -300V, we preferred to add a further -300V and use the neon chain.

The use of the t.r. cell "keep alive" supply, is interesting, this did not exist in our equipment which is a receiver chain only, but I would like to enquire if it is found necessary to stabilize this supply when used as described. Presumably it would not be, if used for "keep alive" purposes only.

Yours faithfully,

G. W. G. COURT  
Air Department,  
Civil Aviation Branch,  
Wellington,  
New Zealand.

## Recurrence Relation for Inductive Circuit with Non-linear Resistance

DEAR SIR.—The differential equation of a circuit consisting of a d.c. voltage  $E$  connected to an inductance  $L$  in series with a non-linear resistance (for example, a diode) having a current-voltage characteristic,  $i = kv^n$ , is given by

$$L(di/dt) + (i/k)^{1/n} = E \quad (1)$$

$$\text{or } Lknv^{n-1}(dv/dt) + v = E \quad (1a)$$

By separation of variables the solution is

$$t_n(v) = nkL \int_{\frac{E}{v}}^{\frac{v^{n-1}}{E-v}} dv \quad (2)$$

In a recent text<sup>1</sup> it is noted that solutions of equation (2) can be obtained in terms of elementary functions for  $n =$  a positive integer or an odd multiple of  $\frac{1}{2}$ , and details are given for  $n = 2, 3, 3/2$ . The case  $n = 3/2$  has also been analysed by Liebetegger<sup>2</sup>.

We note here that a recurrence relation is readily established between the solution of equation (2) for exponent  $n$  and that for exponent  $n+1$ .

Thus

$$\begin{aligned} t_{n+1}(v) &= (n+1)kL \int_{\frac{E}{v}}^{\frac{v^n}{E-v}} dv \\ &= (n+1)kL \int_{\frac{E}{v}}^{\frac{v^{n-1}(v-E+E)}{E-v}} dv \\ &= (n+1)kL \left[ E \int_{\frac{E}{v}}^{\frac{v^{n-1}}{E-v}} dv - \int_{\frac{E}{v}}^{\frac{v^{n-1}}{E-v}} dv \right] \\ \text{so that } t_{n+1}(v) &= \frac{n+1}{n} (Et_n(v) - kLv^n) \quad (3) \end{aligned}$$

Hence if the functions  $t_n(v)$  are evaluated in the range  $0 < n \leq 1$ , either by elementary functions or, where necessary, by numerical methods, their values for  $n > 1$  can be found from equation (3) without further recourse to the defining integrals.

For example, from equation (2)

$$t_1(v) = kL \int_{\frac{E}{v}}^{\frac{v^{-1}}{E-v}} dv = -kL \log(1-(v/E))$$

Then by equation (3), for a square-law device ( $n = 2$ )

$$\begin{aligned} t_2(v) &= 2Et_1(v) - 2kLv \\ &= -2kLE \log(1-(v/E)) + (v/E). \end{aligned}$$

As a second example

$$\begin{aligned} t_{1/2}(v) &= \frac{1}{2}kL \int_{\frac{E}{v}}^{\frac{v^{-\frac{1}{2}}}{E-v}} dv \\ &= \frac{1}{2}kLE^{-\frac{1}{2}} \log \frac{1+(v/E)^{\frac{1}{2}}}{1-(v/E)^{\frac{1}{2}}} \end{aligned}$$

The solution for a  $3/2$ -power law device is then

$$\begin{aligned} t_{3/2}(v) &= 3Et_{1/2}(v) - 3kLv^{\frac{1}{2}} \\ &= 3/2.kLE^{\frac{1}{2}} \log \left[ \frac{1+(v/E)^{\frac{1}{2}}}{1-(v/E)^{\frac{1}{2}}} \right] - 3kLv^{\frac{1}{2}} \end{aligned}$$

Yours faithfully,

H. KAUFMAN,  
Mathematics Department,  
McGill University,  
Montreal.

## REFERENCES

1. ELECTRICAL ENGINEERING STAFF, MIT, Electric Circuits, pp. 687-690 (John Wiley & Sons, Inc., New York, 1940).
2. LIEBETEGGER, A., Circuits with Non-Linear Resistance, *Wireless Engr.* 33, 24 (1956).

## PUBLICATIONS

### RECEIVED

NOTES ON APPLIED SCIENCE No. 13, STANDARD CAPACITORS AND THEIR ACCURACY IN PRACTICE is the title of a recent publication by the National Physical Laboratory. The booklet is the result of many years of experience in testing and using such instruments at the Laboratory. Its purpose is to draw attention to the various factors which affect the performance of capacitors by virtue of their construction and method of use. Primarily the booklet is concerned with capacitors designed for use as laboratory standards of capacitance. Her Majesty's Stationery Office, Kingsway, London, W.C.2. Price 1s. 3d.

GERMAINUM DIODES by S. D. Boon is a book in which the author endeavours to include only sufficient theory to enable the reader to understand the principles of the germanium diode to analyze the most commonly occurring diode problems and to design diode circuits to meet his own requirements. Philips Technical Library, Eindhoven, Holland.

THE RADIO AMATEUR'S HANDBOOK 33rd edition, 1956, contains an enlarged basic theory section and includes a new chapter on semiconductors. Other additions are a series of simple converters to be used with a standard communications receiver, and condensed design data as well as practical construction information on v.h.f. beam antennas. As in previous years, the handbook contains a large catalogue section, featuring communications equipment of the United States' leading manufacturers. The American Radio Relay League, West Hartford, Connecticut, U.S.A. Price \$3.00 in the United States, \$3.50 in U.S. Possession and Canada, \$4.00 elsewhere.

## NORTH AMERICA

### M.I.T.-Harvard Synchrotron

The Massachusetts Institute of Technology and Harvard University will jointly design, build, and operate in Cambridge a six billion volt electron synchrotron to be called the Cambridge Electron Accelerator.

It will be devoted to basic research in the structure of matter. The work of the accelerator will be of general and fundamental scientific interest.

The new "machine" will be built at a cost of about \$6.5m with funds provided by the U.S. Atomic Energy Commission.

Following formal approval of an agreement between the Executive Committee of the Corporation of M.I.T. and the Corporation of Harvard University, the two institutions have formed a joint management committee to operate the project. Work on the detailed plans will begin immediately, and the machine will be completed in about four years. Dr. M. Stanley Livingston, Professor of Physics at M.I.T., will be the first director.

Design studies for a large electron accelerator have been under way by a group from the Physics Departments at Harvard and M.I.T. for the past two years, supported in part by a joint programme of the Office of Naval Research and the Atomic Energy Commission. On the basis of these studies, present plans for the Cambridge Electron Accelerator call for a machine which will accelerate electrons around a circular path 236 feet in diameter. It will be installed in a circular underground trench building with adjacent laboratories for experimental work. The tentatively selected site is on Harvard property adjacent to its Cyclotron.

Electrons entering the circular path at low energy will be accelerated around it by 6 radio frequency circuits. Forty-eight laminated magnets, each measuring 2 by 2 by 11 feet, will guide the electrons round the orbit. The electrons will emerge with very nearly the velocity of light. The increase in the electrons' energy in the machine will be accompanied, in accordance with the Einstein law for the equivalence of mass and energy, by a very large increase of mass; the emerging accelerated electrons will have a mass about 12 000 times that of the entering electrons.

Though the machine to be built in Cambridge will be, so far as is known, the highest energy accelerator of electrons, it will not be as powerful as the proton synchrotron now being built at the Brookhaven National Laboratory (Upton, Long Island, New York), which will accelerate protons to energies of 25 to 30 billion electron volts.

One-billion-volt accelerators of elec-

trons are in operation at Cornell University (Ithaca, New York) and the California Institute of Technology (Pasadena),

### Banks' Computer Project

Agreement between Bank of America and General Electric Company's Electronics Division, Syracuse, N.Y., for the manufacture of ERMA electronic computers for the bank's long-range multi-million-dollar computer programme, was announced recently by the bank's president.

Manufacture of ERMA equipment will be a major addition to G.E.'s Industrial Computer Section programme.

According to the bank's president, Bank of America's contract represents the largest civilian order in history for industrial computers and marks a revolutionary step forward in the banking profession.

Initial delivery of ERMA components is scheduled for next year, and where practicable transistors will be used in the new computers.

### Colour Television

General Electric have announced that they will enter the colour television receiver market in the second half of this year. Details of the receivers have not been given but it is stated that they will use a three-gun aperture mask cathode-ray tube.

### High Purity Solder Microforms

To meet the need for near-absolute purity in solders used to join the parts of transistors, diodes and other semiconductors, an unusual series of solder microforms with rare metal elements that range in purity to 99.999 plus, has been introduced by the Anchor Metal Co. They are available in the following elements: indium, gallium, selenium, tellurium, gold, platinum, silver, aluminium, arsenic, antimony, bismuth, cadmium, copper, tin, zinc and lead—with degrees of purity between 99.9 to 99.999 plus.

The solder microforms can be obtained with just one of these elements or in a wide selection of "doped" alloys such as indium-gallium, indium-gold, indium-tin, gold-gallium, gold-antimony or tin-arsenic. Almost any desired combination can be produced to satisfy design requirements.

Present forms of the solder materials include crystals, cast cake, continuous cast rod, extruded stick, cast stick, pellets, dots, washers or wire.

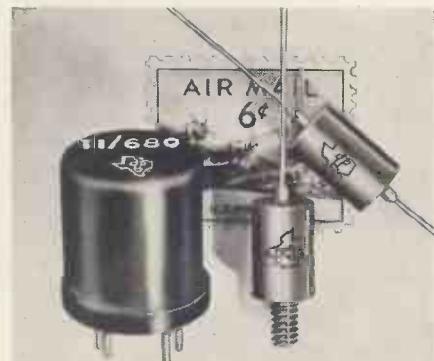
### New Transistor Production Technique

Texas Instruments have recently announced that they are now producing

transistors by what they describe as "grown-diffused" technique which is based on the principles of solid state diffusion, wherein an ultra thin base layer is produced by the migration of added impurity atoms while the semiconductor crystal is being grown.

They state that silicon transistors now in production by this new process have gains of 15dB at 12.5Mc/s and 14dB at 30Mc/s with useful power levels up to 150°C. A "grown-diffused" germanium transistor with a power gain of 10dB at 100Mc/s and with a collector dissipation of 25mW at 75°C is also stated to be in production.

Also announced by the same firm are five new silicon rectifiers, the main feature of these being a single grown junction



A group of single junction rectifiers

element with a 1.5kV breakdown voltage and forward current ratings up to 100mA. These are available in several types, one of which is a plug-in version to replace the 6X4.

### L-Band R.F. Dummy Load and Calorimeter

Levinthal Electronic Products have recently introduced a completely self-contained r.f. dummy load and calorimeter which is designed for applications to high peak powers or high average power installations. As an r.f. dummy load, the unit is capable of handling in excess of 2MW of peak power. The closed circuit cooling system with its pump and heat exchanger permits satisfactory cooling at greater than 10kW of average power.

For calorimetry, the unit uses a substitution method for greatest accuracy. A completely separate L-band waveguide, with a waterload tube containing a 60c/s Nichrome heater wire is used as a calibrating load. A controllable amount of 60c/s power, metered by a 1 per cent wattmeter, can be fed to this calibrating load by means of an externally-operated Powerstat.

The flow rate is controlled by a flow regulator and metered by a flowmeter. Thermopiles across the input and output water, measure the temperature differential and actuate a millivoltmeter which is calibrated to read directly in r.f. watts.