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Commentary

THE transistor has now been with us for ten years, for it was in August 1948 that its invention was first announced in the *Bell Laboratories Record*.

While there are frequently rival claimants to an invention of any significance, there are few who will deny that the credit for the initial invention of the transistor belongs to the Bell Telephone Laboratories and, indeed, the scientific importance of this work by J. Bardeen, W. H. Brattain and W. Shockley was recognized by the award of the 1956 Nobel Prize in Physics.

In the ten years since its invention the progress in the technology of the transistor and its application, and in the whole field of semiconductor physics, has been enormous.

Prior to World War II only a few people had ever seen a piece of germanium and it received its first real attention in the wartime scramble to develop better materials for microwave diodes. By the end of the war polycrystalline germanium of high purity (by the standards of the day) was available and it was during a basic research study of its surface properties, at the Bell Laboratories, that the transistor was discovered.

In the intervening years semiconductor research has continued to increase and it is now a major scientific activity throughout the world. Almost all the effort has been concerned with germanium and silicon, in that order. This, in some respects is a little surprising since, by a conservative estimate, there are at least a thousand semiconductive substances. It is, however, fortunate that transistor action was first observed on germanium as it is one of the simplest of all solids, a factor which has contributed to the fact that it is, today, probably the most completely understood of all solid materials.

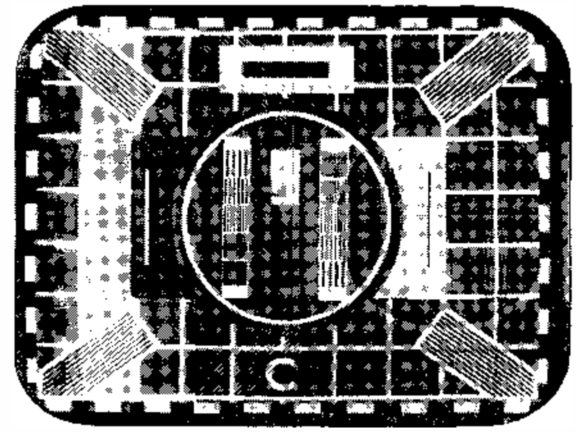
Both germanium and silicon are now available as the purest and most perfectly formed crystals which scientists have had for their study and, even more important, this purity and structural perfection can be altered in a controlled manner by a number of techniques. It is the electrical effects of these alterations that are the distinguishing features of semiconductors. Some of the techniques evolved for dealing with germanium and silicon have an importance which transcends the semiconductor field. The chief of these being zone refining and its variant 'floating-zone refining', which was developed mainly for use with silicon, both of which are major metallurgical inventions.

Research into the basic properties of semiconductors has been backed up by sound progress in the manufacturing techniques necessary to make reliable transistors. To the production engineer the original point-contact transistor appeared as a product simple in structure, rugged and enclosed in plastic so that it would lend itself easily to mass production. Unfortunately, it did not turn out like that, for problems were soon encountered in terms of reliability, reproducibility and range of performance. The advent of the grown junction and alloying and diffusion techniques have also posed their considerable problems to the production engineer who has had to strike a balance between the present —to provide adequate facilities for existing techniques—and the future—to be prepared for the new and better way in this new and rapidly evolving field. Nevertheless, in ten years the transistor has been developed to the point where it can compete both in performance and cost, in many applications, with its far older competitor, the thermionic valve.

With regard to the application of the transistor, while some of the early forecasts that it would almost entirely replace the thermionic valve have proved to be rather wildly optimistic, its impact has been significant and there is now a very large number of transistors in satisfactory use in a very diverse range of electronic equipment. The initial phase where a designer or manufacturer used transistors merely to be fashionable has, fortunately, largely passed and the transistor is being used in applications where its virtues of reliability, small size, low power consumption and low operating voltages, make it a more suitable circuit element than the corresponding thermionic valve.

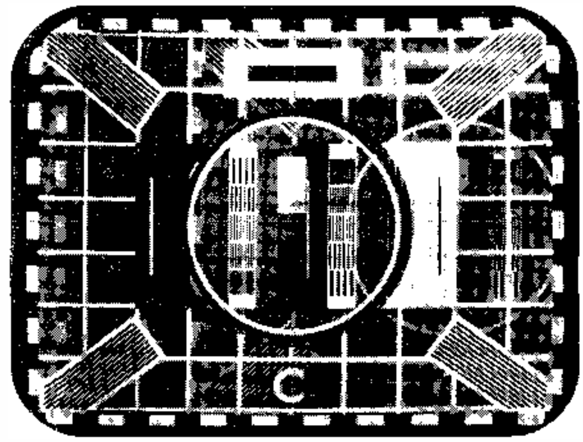
The future of the transistor is difficult to forecast. It is, however, safe to say that as the basic knowledge of semiconductors expands and new and improved manufacturing techniques become available the operating range and reliability and consequently the field of application, will increase in step. The form of the actual transistor is equally hard to forecast but it seems probable that for many applications germanium will be superseded by, perhaps, silicon, which has an operating temperature range up to about 150°C as against some 50°C for germanium; also, silicon is far more plentiful although, at present, the extra difficulties of refining make it rather more costly.

But, whichever way future developments may take us, it is certain that the transistor will rank as one of the great inventions of the past decade.



An Aircraft Simulator for Television Signals

By M. C. Gander*, and P. L. Mothersole*, A.M.Brit.I.R.E.



The instrument described enables the reflected signal from an aircraft to be simulated. The delay, amplitude and rate-of-change of phase of the delayed signal are variable and enable multi-path effects to be studied in the laboratory. A variable mercury delay line is used to obtain the signal delay. An important feature of the instrument is the ability to produce a constantly changing phase between the two signals by other than mechanical means.

The above photographs show a delayed signal with the phase adjusted to give a negative and a positive ghost.

TELEVISION signals are reflected from aircraft. If the reflected signal is received at appreciable strength in comparison to the direct signal a ghost picture is observed. The position and phase of the reflected or ghost signal depends on the path difference between the two signals. Since the aircraft is moving the position and phase of the reflected signal is constantly changing; this gives rise to a beat effect, or 'flutter', on the receiver's screen.

Circuits are often used in a receiver to minimize this effect (e.g. a.c. coupling, a.g.c. etc.) A receiver can normally only be tested by utilizing reflections from aircraft flying in the vicinity of the laboratory. It is therefore difficult to repeat experiments, since the exact reflection is not repeatable at will.

When a new system (e.g. colour television), is being developed in the laboratory it is very important to study the effects of reflected signals on the received picture. Since real aircraft cannot be used the reflected signal must be produced by artificial means.

The aircraft simulator described produces the exact effect of an aircraft. The delay, amplitude and rate-of-change of phase of the reflected signal is variable over wide limits.

A Simple Flutter Simulator

A method often employed to test a.g.c. systems is to vary the amplitude of the r.f. signal input to a receiver, and to measure the resulting change in the video signal amplitude. If the r.f. signal amplitude is varied by a low frequency oscillator a form of flutter is produced at the receiver's output.

This method is satisfactory for a production test, but it cannot produce the true effect of an aircraft necessary for a close study of the problem. An aircraft will reflect a signal which at the receiver will be shifted in phase with respect to the direct signal, and added to it. Thus video information may be superimposed on the synchronizing signals. This effect is of particular interest with a colour signal of the N.T.S.C. type.

A simulator must therefore delay the input signal and produce an output signal that is a summation of the two.

Wide Band Signal Delay

The Mercury Delay Line

The signal delay required by a simulator must be continuously variable up to at least 100μsec and have a bandwidth of some 4Mc/s. A device that is now available to meet this requirement is the variable mercury delay line. The type developed at the Mullard Research Laboratories consists of a mercury filled steel tank some 15in long fitted at one end with two quartz crystal electro-acoustic transducers (Fig. 1). Inside the tank is a sliding piston, driven by a precision lead screw.

The piston is machined to form a corner reflector. The acoustic signal from the input transducer is reflected by the corner reflector, back to the output transducer. Movement of the reflector therefore varies the path length in the mercury, and hence the time delay. The total delay in a line

* *Mullard Research Laboratories.*

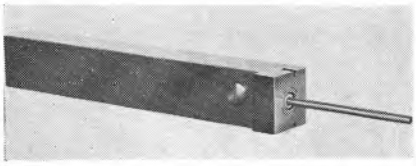


Fig. 1. Ultrasonic mercury delay line type YL2100

of this type is about 300μsec. The lead screw is normally cut with the exact pitch required to make one revolution correspond to 10μsec change of delay time.

The frequency response of a mercury delay line is governed by the characteristics of the quartz crystal units. A frequency response of some 0·86 times the band centre (−3dB), may be obtained. The insertion loss of such a line is about 50dB at 15Mc/s, into a terminating impedance of 300Ω.

The linearity of a mercury line is very wide, extending from very small signals to signals of a few hundred volts applied to the transmitter crystal.

Video Signal Delay

One method of producing the signal to test a television receiver is to modulate the video signal on to a carrier frequency of about 15Mc/s (Fig. 2(a)). The modulated signal is then amplified and applied to the line. The output from the line will then require further amplification and demodulation.

It is very important to ensure that the sync-to-picture ratio of the output signal is correct, and that the black level of the delayed signal follows that of the input signal.

It is also important to ensure that the signal delay does not introduce undue noise. Noise produced by the first stage of the output amplifier will be the main limitation, and the signal level at this point must be about 3mV to achieve an overall signal-to-noise ratio of about 50dB. The transmitting crystal therefore requires a drive of about 1V.

R.F. Signal Delay

An alternative method of producing a signal to test a television receiver if the signal is available at r.f., is to frequency change the carrier to about 15Mc/s (Fig. 2(b)). Since it is desirable to have the output signal at the same frequency as the input, a second frequency changer is required after the line. A common local oscillator may be used for both frequency changers which ensures that the input and output signals have the same frequency.

Complete Simulator

A block diagram of the complete simulator is shown in Fig. 3.

Flutter Generation

As the video signal was available at r.f. it was decided to use the r.f. delay method described above to delay the signal. In a simulator the delayed signal must be added to the non-delayed signal and to produce flutter the phase (or time delay) of the delayed signal must be varied.

A convenient method of adding the two signals is to bridge the mercury line with a variable attenuator. Since the line produces about 50dB attenuation the direct signal amplitude may be made larger or smaller as desired.

One method of producing a constantly changing phase (or time delay), is to move the reflector in the mercury line with a small motor. If a variable attenuator is ganged to the motor shaft the amplitude of the reflected signal may be varied automatically as the time delay varies. Thus the exact effect of an aircraft moving close to a receiver's aerial would be simulated.

When studying the effects of a reflected signal in a receiver it would be more convenient to position the reflected signal at some required fixed time delay, to independently adjust its amplitude, and to cause its phase

Fig. 2. (a) Video signal delay (b) R.F. signal delay

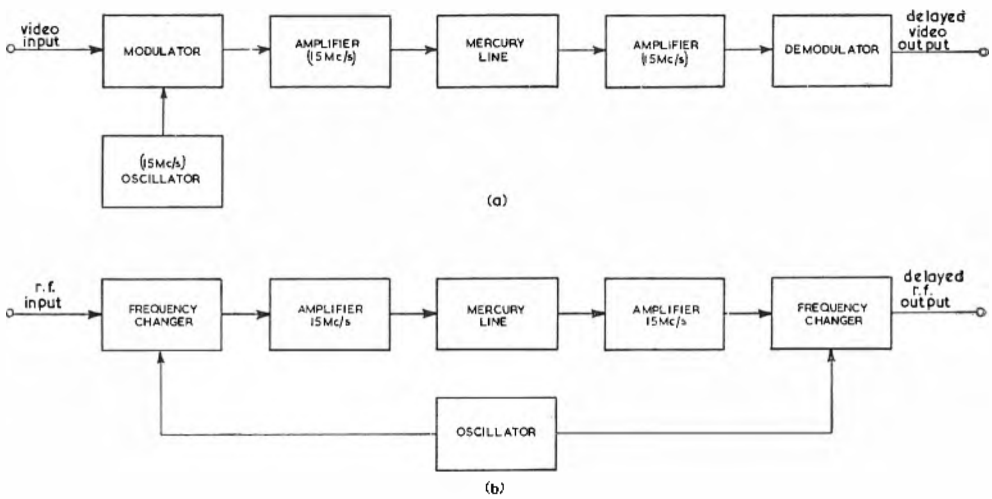
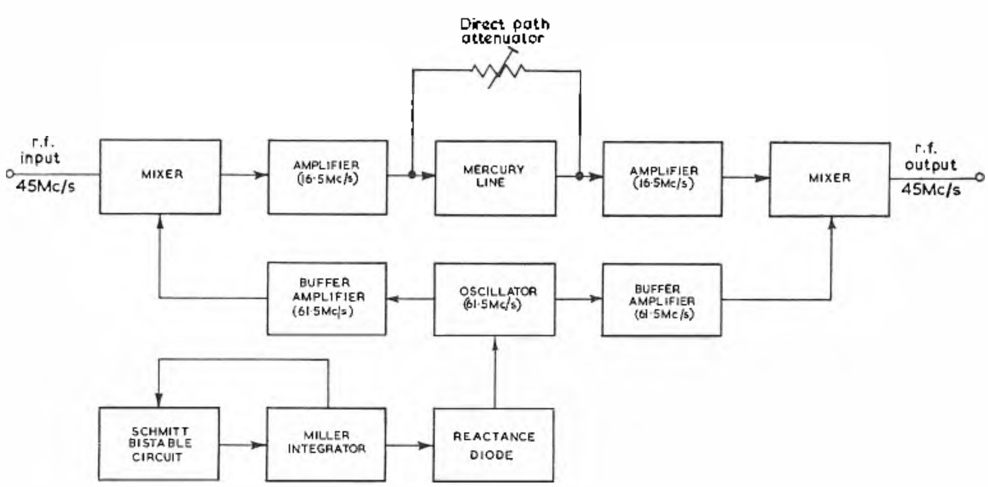


Fig. 3. Aircraft simulator block diagram



to rotate at a constant rate. This would be equivalent to a real aircraft giving a reflection whose amplitude was adjustable, at the same time the aircraft would be oscillating on the spot a distance of one wavelength. In practice this is impossible but this effect may be produced quite simply in the simulator.

The time delay produced by the mercury line is independent of the carrier frequency; the phase between the direct and delayed signal, however, is proportional to the carrier frequency. Thus if the oscillator frequency is varied the signals will vary in phase; the phase may rotate a full 360°. The rate-of-change of phase is proportional to the frequency modulation of the carrier frequency.

The variable mercury line is thus used to determine the time delay (i.e. position of the 'ghost' signal); the shunt

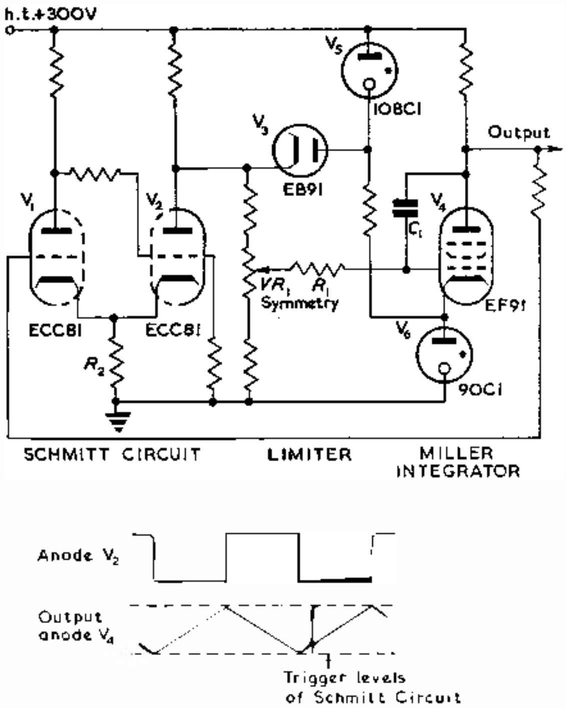


Fig. 4. Sweep generator circuit (simplified)

attenuator the amplitude of the non-delayed signal; and the sweep oscillator frequency the rate-of-change of phase of the two signals (i.e. flutter frequency).

Sweep Generator

To produce a constant rate-of-change of phase a linear change of oscillator frequency is required. The ideal arrangement would be for the swept oscillator to produce exactly 360° phase change per cycle of sweep. In practice, however, this is not simply arranged since the phase shift is proportional to the time delay in the mercury line and also the oscillator frequency. The solution adopted was to frequency modulate the local oscillator by a large amplitude sawtooth wave, producing a phase shift corresponding to many cycles across the line for each cycle of the sawtooth. The discontinuity that occurs when the rate-of-change of the sweep potential changes sign is then insignificant.

A simplified diagram of the sweep generator is shown in Fig. 4. It consists of a Miller integrator d.c. back coupled to a Schmitt bistable circuit. The operation is as follows.

Assume V_1 conducting and V_2 cut-off. The Miller valve's grid aiming potential will be positive, and the anode potential will therefore be falling linearly, the rate being determined by R_1C_1. As the Miller anode potential falls the potential at the grid of V_1 will also fall, reducing the current flowing in V_1. When the trigger point of the Schmitt circuit is reached V_2 conducts, cutting off V_1. The Miller valve's grid aiming potential is now negative and its anode will therefore rise linearly. Due to the hysteresis effect of the Schmitt circuit it will not trigger back to its original state until the grid potential of V_1 has risen to a positive value determined by the current in V_2 and the cathode resistor, R_2. Providing that the positive and negative aiming potentials for the Miller integrator are equal in magnitude a symmetrical sawtooth will be produced. The aiming potential is adjustable (VR_1) and the limiting diode V_3 and neon stabilizer V_5 ensure a constant amplitude switching waveform. The symmetry of the output waveform is thus independent of the Schmitt circuit.

An important feature of the circuit is that it is d.c. coupled. The amplitude of the output waveform is therefore constant for all frequencies, the frequency being determined by the Miller valve's grid time-constant only (R_1C_1).

R.F. Oscillator Frequency

The simulator was designed for operation on Channel 1 (carrier frequency 45Mc/s). Operation of the mercury line with certain carrier frequencies would give rise to harmonics in the output producing patterning effects on the final picture. The i.f. frequency used was therefore about 16·5Mc/s.

The required local oscillator frequency is therefore about 61·5Mc/s. Since all the amplifiers are wide band the exact frequency of the local oscillator is not important. It is, however, important to ensure that the oscillator does not produce spurious changes in frequency, due, for example, to a microphonic valve. The oscillator valve was therefore mounted on an anti-microphonic base and a semiconductor diode used as a reactance device. The effective capacitance of a semiconductor diode, biased in the reverse direction, is a function of the bias potential. If a diode is therefore connected into an oscillator valve's tuned circuit, variation of the bias potential will vary the oscillator frequency. Such a device has the advantage of being completely free from microphonic effects.

Signal Levels

The simulator was designed to operate with an r.f. input potential of about 1mV. The gain of the input frequency changer and i.f. amplifier is about 60dB, providing the line with a drive of about 1V. The output i.f. amplifier and mixer provide an r.f. output signal of about 15mV.

Complete Circuit Description

Input Circuits

The input frequency changer and i.f. amplifier circuit is shown in Fig. 5. The 45Mc/s input signal is coupled to the control grid of frequency changer valve V_1 by a matching transformer T_1. The local oscillator voltage (61·5Mc/s) is applied to the cathode via the transformer T_3. The difference frequencies (16·5Mc/s vision and 20Mc/s sound) are coupled through transformer T_3 to the input i.f. amplifier.

The i.f. amplifier is a four stage stagger-tuned amplifier, with a cathode-follower output (V_6). The cathode-follower drives the transmitting transducer of the mercury line directly and also the direct path attenuator, via a resistive divider (R_{21}, R_{22}).

Output Circuits

The output frequency changer and i.f. amplifier circuit is shown in Fig. 6. The mercury line is connected directly

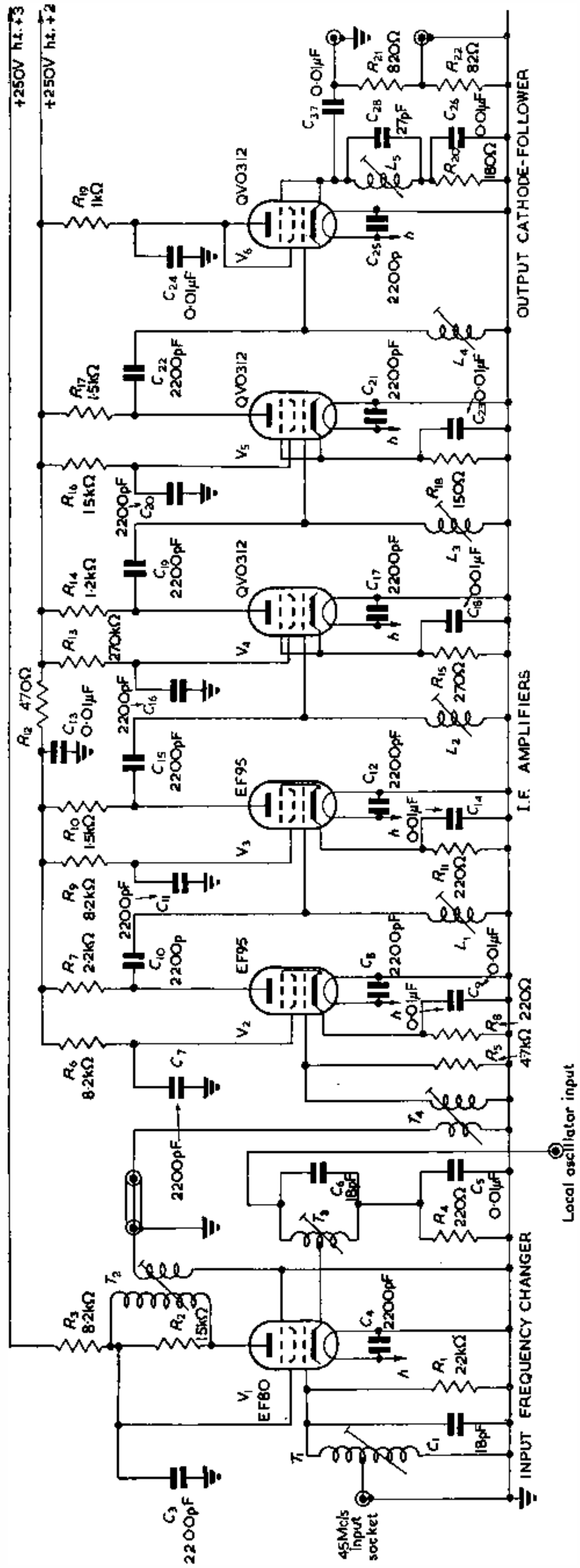


Fig. 5. Input frequency changer and i.f. amplifier

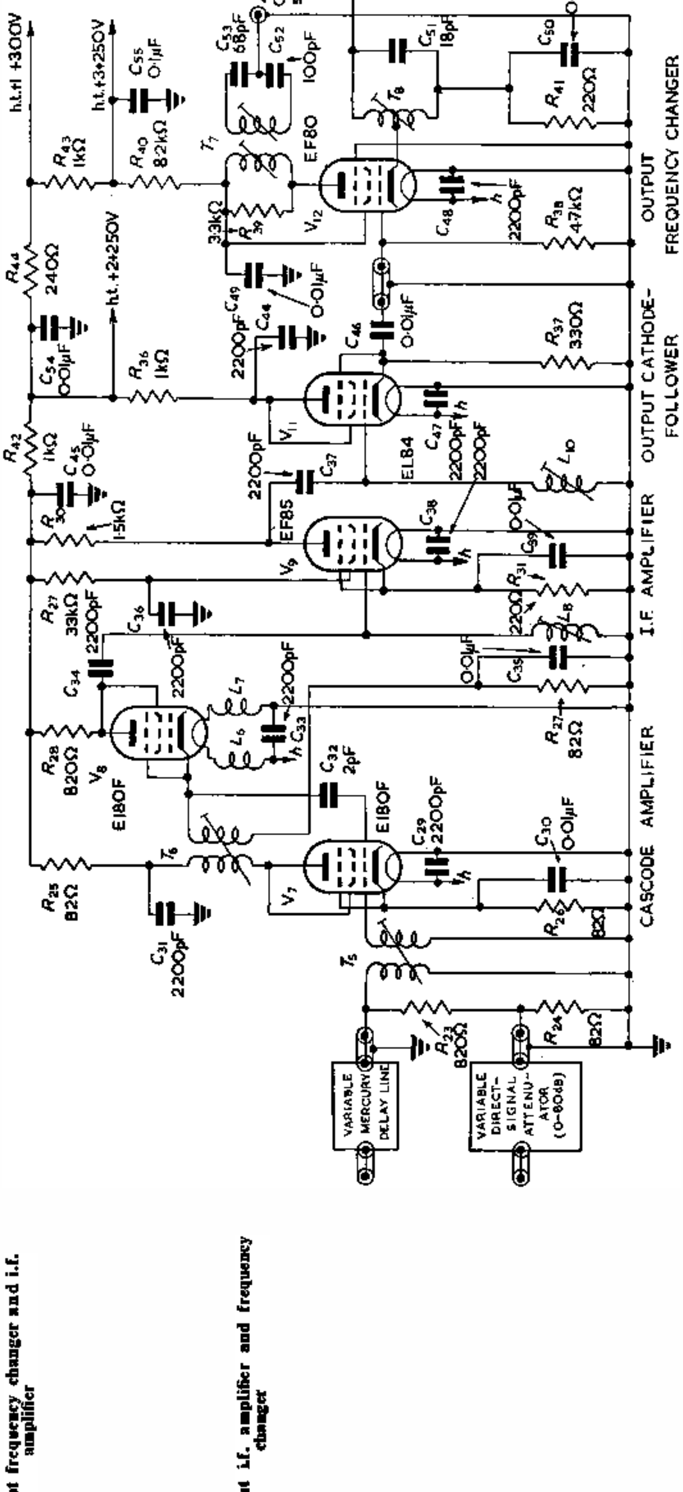


Fig. 6. Output i.f. amplifier and frequency changer

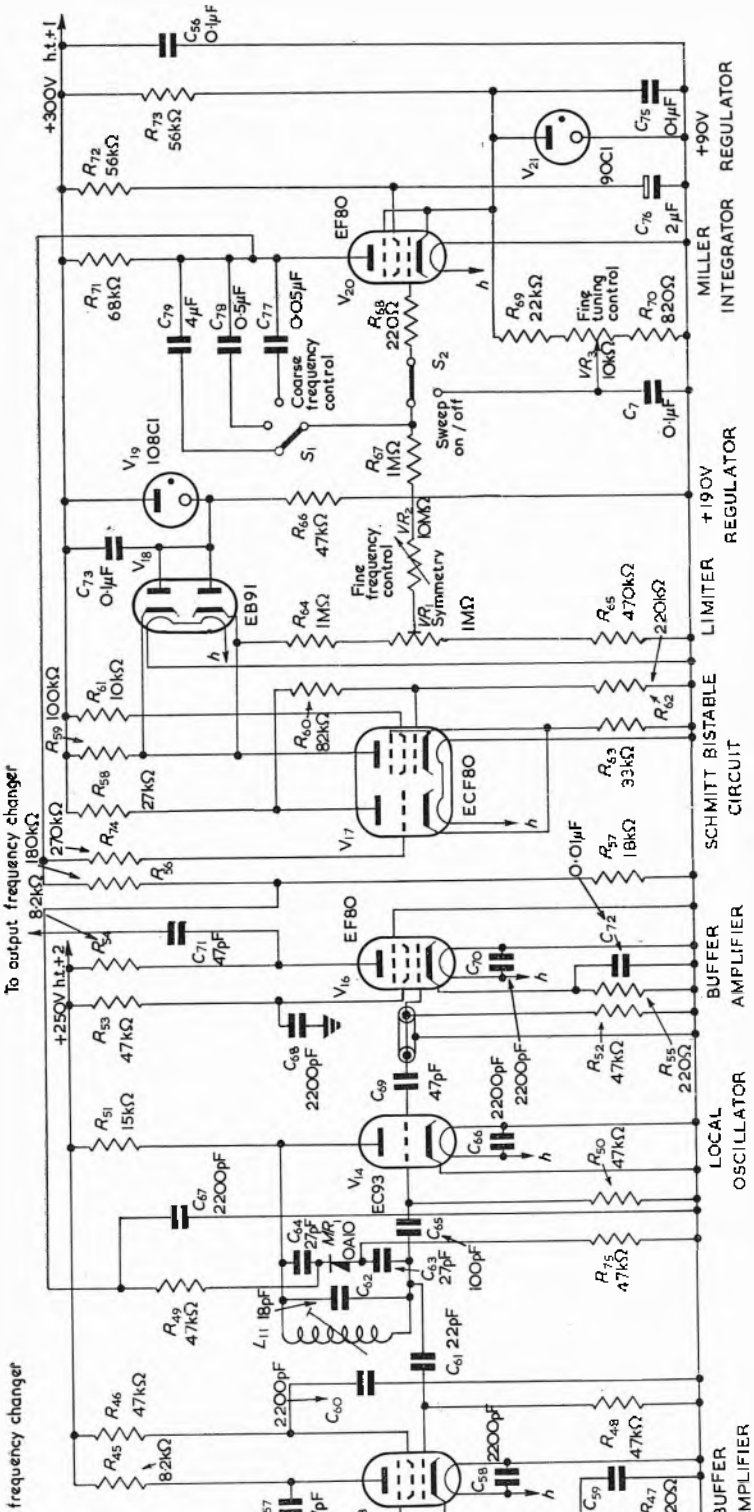


Fig. 7. R.F. oscillator and sweep generator

to the input transformer T_5, the direct path attenuator being connected via a resistive divider (R_{23}, R_{24}). To obtain the best signal-to-noise ratio the first amplifier stage consists of a cascode amplifier, using two high slop triode connected pentodes (V_7, V_8, E180F). The following stage is a tuned amplifier driving a cathode-follower output stage (V_{11}).

The output frequency changer (V_{12}) has the local oscillator voltage applied to its cathode via the transformer T_5. Signals from the i.f. amplifier are applied directly to the control grid. The difference frequencies (45Mc/s vision, 41·5Mc/s sound), are coupled via the band-pass transformer T_7 and impedance matching divider (C_{52}, C_{53}), to the output socket.

R.F. Oscillator and Sweep Generator

The local oscillator (V_{14}) is a low microphony triode (EC93) operating as a Colpitt's oscillator at 61·5Mc/s. A semiconductor diode (MR_1) is included in the resonant circuit so that the junction capacitance is effective in determining the oscillator frequency. The diode is biased in the reverse direction by a voltage which may be varied between 12 and 18V. This causes a change in junction capacitance which produces a change in the oscillator frequency of approximately 300kc/s.

The buffer amplifiers (V_{13}, V_{16}), prevent the direct leakage of the r.f. input signal between the input and output frequency changers via the local oscillator.

The sweep waveform is produced by the Miller integrator (V_{20}) and the bistable Schmitt circuit (V_{17}). The linear output from the integrator (V_{20}) is applied to the reactance diode (MR_1) via a potential divider (R_{56}, R_{57}).

The bistable Schmitt circuit is controlled by the potential at the grid of V_{17a} which is d.c. connected to the Miller valve's anode via R_{74}. The positive-going Miller waveform carries the grid of V_{17a} positive until V_{17a} conducts cutting off the pentode V_{17b}. The anode potential of V_{17b} rises to h.t. cutting off the limiter diode V_{18}. The rise in potential also lifts the aiming potential of the Miller valve positive, causing a linear fall in potential at the Miller valve's anode. The Miller anode potential will then continue to fall, taking the grid of V_{17a} negative, until V_{17b} conducts sufficiently to reverse the Schmitt circuit. At this instant V_{17b} anode falls to 190V where it is held by the diode V_{18}. The aiming potential is now again negative (with respect to the cathode of the Miller valve), causing the Miller anode potential to rise.

The Miller integrator thus produces a triangular waveform of constant amplitude, 60V peak-to-peak. The frequency is variable between 0·003c/s and 1·3c/s approximately, depending on the setting of S_1 and VR_2 (coarse and fine frequency controls).

The automatic sweep may be switched off by S_3, and the anode potential of the Miller valve set by VR_3 (local oscillator fine frequency control).

The regulator neon V_{21} stabilizes the Miller valve's cathode at 90V to obviate a negative supply for the grid circuit network R_{64}, R_{65}, VR_1. The regulator neon V_{19} and limiting diode V_{18} ensure a constant switch waveform from the Schmitt circuit. The symmetry of the sweep waveform may be adjusted by VR_1 (preset).

All the r.f. circuits are designed to operate from a 250V supply; the sweep circuits from 300V stabilized supply. The total h.t. supply current is 260mA at 300V.

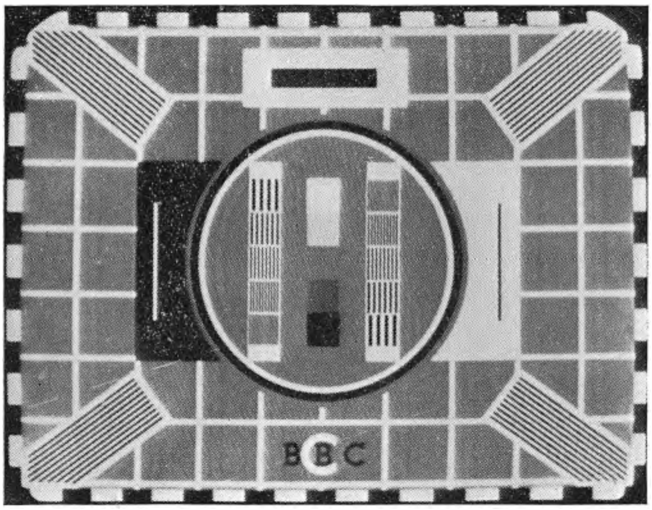


Fig. 8. Signal after a delay of 100μsec

Operation

Time Delay

The minimum delay obtainable in the variable mercury line is about 30μsec and is due to the mechanical construction. To obtain a delay from zero a small fixed delay of about 30μsec could be used in series with the direct path attenuator (Fig. 3). This would complicate the instrument and for many purposes is not necessary. In practice a delay of about 100μsec, corresponding to one line, is quite satisfactory to simulate a reflection with a very small time delay.

Reflection Amplitude

The reflected signal (i.e. delayed signal) amplitude could be adjusted by a variable attenuator in series with the mercury line. However, to obtain a good signal-to-noise ratio the output from the line should be maintained at about 3mV. To maintain this signal level the input to the line must be about 1V, and to provide a drive higher than this is not economic. The alternative method is to use the delayed signal as the 'direct' signal, and to vary the amplitude of non-delayed signal. This enables the combined output signal to have the best signal-to-noise ratio since the signal from the mercury line is not attenuated. The direct path attenuator (Fig. 3), is in parallel with the mercury line and can provide an attenuation of between 20 and 100dB (0 to 80dB variable, 20dB fixed).

Rate-of-Change of Phase

The rate-of-change of phase between the two signals is determined by the sweep oscillator. The frequency is adjustable and produces a flutter frequency between 0·2c/s and 100c/s on a receiver. When the sweep oscillator is switched off the phase may be adjusted by the oscillator fine tuning control (VR_3, Fig. 7).

Results and Conclusions

The use of a variable mercury line to delay a television signal has been described in this article. The results obtained on a receiver with test card 'C' are shown in Figs. 8 and 9. For these photographs external synchronization of the receiver was used. Fig. 8 shows the signal after a delay of 100μsec (1 line). For this exposure the 'direct path' attenuator was at maximum attenuation (100dB). The 'ghost' signal, with about 8μsec delay, is a site reflection. In Fig. 9 the delay has been increased to about 120μsec and the direct signal amplitude adjusted to that of the

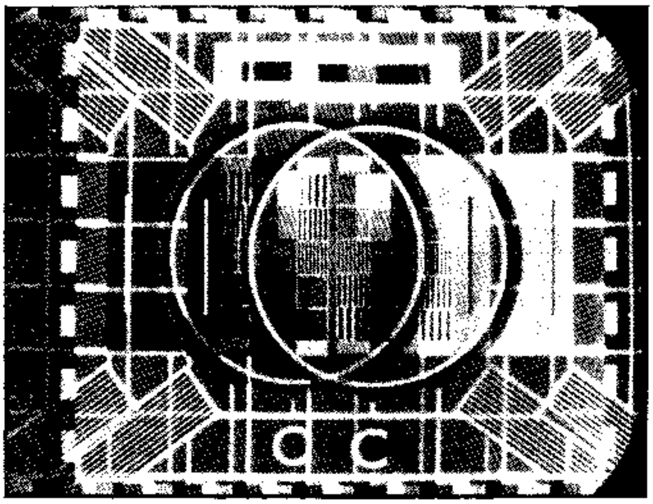


Fig. 9. A delay of 120μsec, amplitude of direct and delayed signal equal

delayed signal. It can be seen from Fig. 9 that the two signals are substantially the same. In the photographs on page 408 delay is about 140μsec and the direct signal amplitude is greater than the delayed signal. The phase has been adjusted (oscillator fine tuning VR_3, Fig. 7) to produce a positive ghost signal on one, and a negative ghost signal, on the other.

The low frequency sweep circuit enables the characteristics of the line to be exploited to obtain a constantly changing phase between the delayed and non-delayed signals, without the use of mechanical devices. The instrument produces the exact effect of an aircraft and thus enables the effects of aircraft on a receiver's performance to be studied in a laboratory. This facility is particularly valuable when appraising a new system, such as the N.T.S.C. system. In addition to its application for television signals the simulator could be very simply adapted to assist in the study of problems associated with f.m. and v.h.f. communications systems.

Acknowledgments

The authors would like to acknowledge the contribution made by Messrs. D. H. Hardy and J. S. Palfreeman in supplying a modified prototype delay line for use in this instrument and also two experimental i.f. amplifiers. They also wish to thank the Director of the Mullard Research Laboratories and the Directors of Mullard Limited for permission to publish this article.

The variable mercury delay line used in the instrument is now manufactured and marketed by Mullard Equipment Limited, under the type No. YL 2100.

Aerial Feeders for Multi-channel Links

By L. Lewin*, A.M.I.E.E., and J. Payne†, B.Sc., A.Inst.P.

The main adverse effect of long feeders in a microwave link is to introduce attenuation and distortion, the latter arising mainly from multiple reflections within the feeder. The reflections from coaxial cable and waveguide feeders are investigated and the requirements for broad-band links explained. The suitability of the various types of feeder for different frequency bands is outlined, high quality waveguide runs being preferred except for narrow band links at the lower frequencies.

In portable radio link systems it is customary to mount the radio frequency head unit adjacent to the aerial. When the propagation requirements of the transmission path make it necessary to use the head-aerial combination high on tower or mast the inaccessibility of the head-unit for maintenance purposes is tolerated since the installation is only temporary.

For permanent systems this condition is unacceptable and it is necessary to install all the radio equipment at ground level with a radio frequency connexion or feeder to the aerial which might need to be as high as 300ft above the ground.

The necessity for a feeder is considerably increased when several equipments working at different frequencies are to make use of a common aerial.

Transmission Requirements

Over the frequency range of each signal channel the insertion loss and phase-shift properties of the feeder must be such as to produce no more than its allocated contribution of thermal noise and distortion to the total thermal noise and distortion of the whole link.

The most severe requirement arises in the case of a f.d.m./f.m. signal for a telephone system with a large number of channels.

As a consequence of multiple reflections of a frequency modulated wave in feeders preceding an ideal discriminator, harmonic and intermodulation distortion will appear at the output. The mechanism of this effect has been described before[1,2] and the formulae and results of this previous work will be drawn on now to obtain the reflection requirements of a feeder.

In the case of the transmission of an N.T.S.C. type colour television signal, the requirements are comparable with those for a 600-channel telephone system. They are much less severe for a monochrome signal.

The total reflection distortion can be resolved into four components:

(a) Due to an echo formed by reflection from the aerial and re-reflection from the radio transmitter; or, in the case of a receiving terminal, by reflection from the input to the radio receiver and re-reflection from the aerial.

(b) Due to the sum of the echoes formed by multiple reflections within the feeder.

(c) Due to the sum of the echoes formed by reflections between the aerial and the impedance discontinuities within the feeder.

(d) Due to the sum of echoes formed by reflections between the equipment at the bottom of the feeder and the impedance discontinuities within the feeder.

Consider the case of the system being tested by loading the base-band with white noise and measuring the noise density (power) ratio in the top channel[3].

This method may be used for system testing, though the system performance will not be identical to the figures obtained on account of the different statistics of white noise as against telephone conversation loading and speech. From Ref. 2 the contribution of feeder distortion to the noise density power ratio is

$$P = (\omega_N^2/4x)[r_E^2 r_A^2 F_1 A_1 + Mr^2(r_E^2 + r_A^2) F_2 A_2 + M^2 r^4 F_3 A_3] \quad \ldots\ldots\ldots\ldots (1)$$

The mean square reflection coefficient looking into a feeder is

$$r_0^2 = r_L^2 A_1 + Mr^2 A_2 \quad \ldots\ldots\ldots\ldots (2)$$

where r_L is a load reflection coefficient (which might be r_E or r_A).

As an example to show the orders of magnitude involved, a 600-channel system with a top video frequency of 2·64Mc/s and r.m.s. deviation of 0·64Mc/s will be considered. It will be assumed that reflections at the ends of the feeder are 3 per cent and the feeder itself has a total attenuation αL of 4dB = 0·46N and within the feeder there are 20 discontinuities of one per cent reflection each, and a group velocity of 0·7 light velocity in a 200ft feeder.

Then Tx = 1·23 and from Fig. 2 of ref. 2

$$F_1 = 0{\cdot}86,\ F_2 = 0{\cdot}48,\ F_3 = 0{\cdot}13$$

From Fig. 3 of Ref. 2 with αL = 0·46, the attenuation factors are

$$A_1 = 0{\cdot}17,\ A_2 = 0{\cdot}47,\ A_3 = 0{\cdot}6$$

The three contributions to the noise density power ratio in equation (1) are 5×10^{-7}, $3{\cdot}4 \times 10^{-6}$ and $1{\cdot}3 \times 10^{-6}$ respectively giving a total contribution of $5{\cdot}2 \times 10^{-6}$ or −53dB.

Such a level of distortion would be unreasonably high from this source alone, even allowing for the fact that not all feeders in a microwave link would be so long and therefore containing so many points of reflection. (Bear in mind, also, that these figures refer to a white noise test, not to link distortion figures under working conditions.) The reduction

Symbols Used

$T = L/V_g$ = group delay for a transit one way along the feeder.
x = r.m.s. angular deviation of total video signal.
ω_N = highest video angular frequency.
αL = total feeder attenuation (one way) in nepers.
F_1, F_2, F_3 = distortion functions associated with the components *a*, *c* and *d*, *b*.
A_1, A_2, A_3 = attenuation functions associated with the components *a*, *c* and *d*, *b*.
r = voltage reflection coefficient (subscripts *E*, *A* for equipment and aerial).
M = number of discontinuities, of reflection coefficient r, within the feeder.

* *Standard Telecommunication Laboratories Ltd.*
† *Standard Telephones & Cables Ltd.*

in distortion obtained by different choices of feeder parameters will now be considered.

A much lower feeder attenuation can be obtained, if needed, by using a larger waveguide, either circular, rectangular or coaxial, in an overmoded condition. Alternatively, a higher attenuation can be obtained by a deliberate insertion of loss or by the use of lossier materials for the cable. Since thermal noise is increased by increasing the feeder loss, while intermodulation noise is decreased, an optimum must exist. If the total feeder loss is increased to 8dB the attenuation functions become

$$A_1 = 0{\cdot}025,\ A_2 = 0{\cdot}26,\ A_3 = 0{\cdot}39$$

The noise contributions in equation (1) are $0{\cdot}08 \times 10^{-6}$, $1{\cdot}92 \times 10^{-6}$ and $0{\cdot}86 \times 10^{-6}$ respectively, giving a total of $2{\cdot}86 \times 10^{-6}$ or $-55{\cdot}5$dB. Thus, compared to the previous case, the distortion is reduced by 2·5dB while the thermal noise is increased by 4dB. At the other extreme, if the cable introduces no loss at all the A functions are all unity and the distortion contributions are $2{\cdot}9 \times 10^{-6}$, $7{\cdot}2 \times 10^{-6}$ and $2{\cdot}2 \times 10^{-6}$, giving a total of $12{\cdot}3 \times 10^{-6}$ or $-49{\cdot}1$dB. The distortion is increased (from the 4dB case) by 3·9dB and the thermal noise decreased by 4dB. The correct operating region depends on the precise allocation of thermal and distortion noise, but it is apparent that the optimum is fairly flat and that feeder attenuation is not necessarily a bad thing. It is also apparent from the above figures that the dominant contribution, for this length of feeder, comes from reflections from the internal discontinuities to the two end terminations, with purely internal reflections a runner-up. It is clear that the reflections within the feeder must be improved: in fact the above example of one per cent reflection was chosen so as to stress the need for a good feeder, equation (2) giving a mean feeder reflection of $r_0 = \sqrt{(20 \times 10^{-4} \times 0{\cdot}47)} = 3{\cdot}1$ per cent with a good load. (A peak reflection of some three times this value would be found over a sufficiently wide frequency band—the positioning of this band is, of course, outside the engineer's control).

If the intra-feeder reflections are 0·5 per cent then, again with a 4dB feeder loss, the distortion contributions are 5×10^{-7}, 8×10^{-7}, $0{\cdot}8 \times 10^{-7}$, giving a total of $1{\cdot}4 \times 10^{-6}$ or $-58{\cdot}6$dB. The end-to-end reflections are now nearly as important as the internal-to-end reflections.

As a further example, if the feeder length is reduced the distortion is still further lowered, partly on account of the reduced values of the F functions, and partly due to the reduction in the number of internal reflecting points. These more than compensate the decrease in attenuation. Thus, taking a 100ft feeder of 2dB loss, $F_1 = 0{\cdot}6$, $F_2 = 0{\cdot}1$, $F_3 = 0{\cdot}02$ and $A_1 = 0{\cdot}4$, $A_2 = 0{\cdot}65$ and $A_3 = 0{\cdot}75$. With ten internal reflection points of 0·5 per cent each, and the same terminations as before one obtains distortion contributions of $8{\cdot}2 \times 10^{-7}$, $1{\cdot}2 \times 10^{-7}$ and 4×10^{-9} giving a total of $9{\cdot}4 \times 10^{-7}$ or $-60{\cdot}3$dB. It is to be noted that the end-to-end effect is now dominant, and this can more properly be regarded as a shortcoming of the terminations rather than of the feeder.

For systems carrying a smaller number of telephone channels it is seen from the formula that the distortion is reduced by 6dB/octave as the frequency of the highest channel ω_N is reduced. (In practice the r.m.s. deviation x may also be reduced).

The limit set by the system designer on feeder distortion will be determined to some extent on the way the total permissible noise and distortion is allocated throughout the system, but it is clear from the foregoing examples that the general requirement is that the reflection coefficient at any discontinuity within the feeder shall be less than about 0·5 per cent with the number of these points of reflection held as low as possible.

The C.C.I.R. has recommended a frequency plan catering for six two-way r.f. broad-band channels within a bandwidth of 400Mc/s. For low capacity systems of 60 to 120 channels, the problem is still under consideration; the bandwidth occupied by a suitable frequency plan may well be as much as 200Mc/s, although it is not likely to exceed this.

The feeder reflection requirements must be met over the whole range of such bands, and where a major link band and a minor link band are adjacent, it is desirable for the feeder system to cover a bandwidth of 600Mc/s.

The C.C.I.R. frequency plan arranges for alternate r.f. carriers to be mutually cross polarized and it is therefore convenient that an initial installation shall consist of a single feeder and aerial which will carry the three transmit and the three receiver r.f. carriers with the same polarization. Under these conditions when the feeder is handling signals with a level difference of about 60 to 80dB it is necessary that the feeder shall behave as a linear network and the two high level transmit frequencies f_1, f_2 do not generate an appreciable amount of frequency $f_3 = 2f_1 - f_2$. Since the frequency allocation is chosen so that harmful non-linear products fall between received signals this requirement is not difficult to meet, but it is a point which must not be overlooked.

At repeater stations the feeders to the two aerials associated with the two directions of transmission may be in close proximity for a considerable portion of their length. It is important that the external coupling between the two feeders (due, for example, to leakage through flanged joints) should be sufficiently low as to cause no appreciable increase of the aerial coupling.

Possible Types of Feeder

The best way of avoiding feeder loss and reflection distortion is to replace the feeder, or most of it, by free space, i.e., to bring the beam forming element to a convenient low level and use a passive reflector at the appropriate height. This scheme is more properly dealt with as an aerial system rather than a type of feeder; suffice it to say here that one of the disadvantages is that it is not capable of the same degree of decoupling as a conventionally fed aerial.

Although open multi-conductor circuits are unsuitable at u.h.f. and s.h.f. due to radiation effects, during the last ten years circuits have again aroused interest at these high frequencies when used to guide surface modes. The G-line or guide-line consists of a single conductor coated either with a thin dielectric layer or, in some cases, a rather thicker layer of expanded dielectric. The purpose of the dielectric is to slow the wave down by a small amount: associated with this slowing down is an approximately exponential radial diminution of field so that, beyond a certain radius the field is effectively zero. The system therefore differs from a coaxial line in that an outer conductor for a return current is not needed; as a consequence the question of keeping the inner conductor concentric does not arise.

Although the attenuation of a G-line can be kept low enough it does not appear to be a popular medium for feeders. In part this is due to its susceptibility to weather, rain drops on the surface causing both increased losses and reflection. Departure from straightness can cause both radiation and reflection, while the matching at the far ends, involving entry into conical horns, can give rise to radiation pick-up at a level which, though low, may yet not meet the severe requirements for a microwave link. Insertion into a rigid pipe may overcome both the effects of rain and high winds, but due, possibly, to lack of sufficient data, this type of feeder has not found favour for permanent installation in high grade links.

Coaxial cable avoids the disadvantages of radiation by completely enclosing the field, and since it can be economically produced in long lengths and is simple and convenient to install, it has been widely used for aerial feeders. Impedance discontinuities associated with spacer disks have been overcome by positioning the inner conductor by a continuous helical membrane of Polythene or Polystyrene tape and such discontinuities can be kept below one per cent at any particular point along the cable.

Below 1 000Mc/s 'air-spaced' coaxial feeders are used exclusively; above 3 000Mc/s, waveguide is similarly unchallenged. In the intervening part of the spectrum the 2 000Mc/s communication band, stretching from 1 700 to 2 300Mc/s, is widely used for microwave links carrying television or multi-channel telephony.

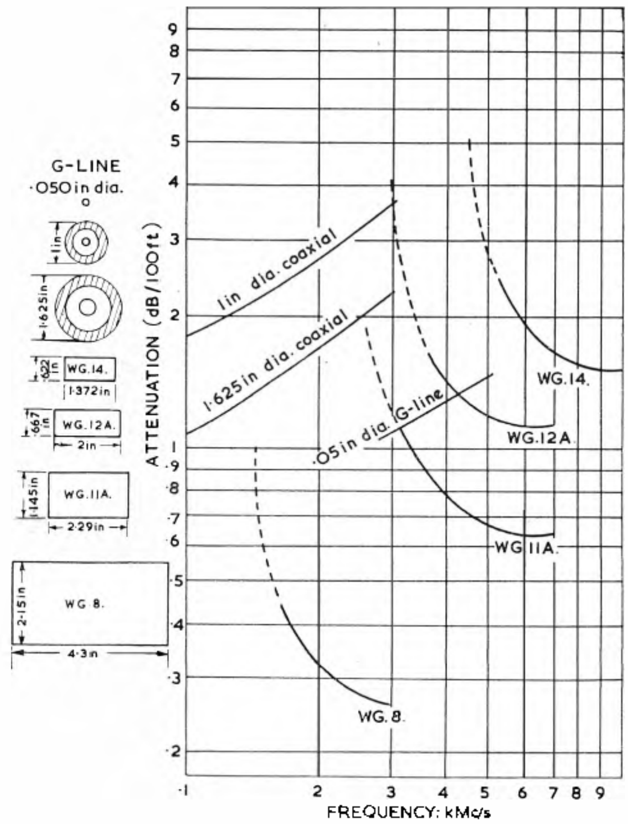


Fig. 1. Comparative attenuation of coaxial and copper waveguide feeders

'Air-spaced' coaxial feeder with an aluminium outer conductor and copper inner conductor supported by a helical membrane, which represents about 10 per cent of the volume between the two conductors, is widely used. The membrane is cut or moulded from Polythene. The main advantages of the coaxial feeder are cheapness, lightness and ease of installation; it is also non-dispersive. It is manufactured in long lengths and wound on drums in much the same way as power cables. The length required for a particular installation is cut in the factory and plugs and sockets fitted. After transporting to the site on a drum it is hauled into position as a single length. The cable is very light, 100ft of 1in nominal diameter weighing under 40lb (depending on the serving), and the same length of $1\frac{5}{8}$in diameter 86lb; for comparison 100ft of aluminium waveguide with flanges every 12ft weighs about 160lb. (The approximate costs are in the ratio 1:2:20). The minimum bending radius of the coaxial feeder is quite small, say 12in to 18in but care is needed during installation to avoid kinking.

The final alignment of aerials can be made before the last cable cleats are fitted, the cable being sufficiently flexible. In general, the cable is bent to suit the contours of the tower, but where necessary a simple cable tray is used for supporting.

There are a number of disadvantages associated with the use of coaxial cable, particularly when large capacity systems are required. The main difficulty is in obtaining a perfectly uniform dielectric support, as irregularities cause internal reflections. An r.m.s. reflection of 5 per cent and a peak reflection of about 10 per cent seem typical figures for a matched 100ft coaxial cable, though an r.m.s. as low as 3 per cent and a peak as high as 20 per cent have been met on some cable systems. The match into the antenna is typically 3 to 5 per cent. The attenuation is higher than that for waveguide designed for the same frequency band, though, as already discussed, this is not necessarily a disadvantage, especially if internal reflections cause signal distortion at a level comparable to that from thermal noise. A comparison of coaxial cable and waveguide theoretical attenuations in the 2 000Mc/s band is given in Fig. 1. The actual attenuations will depend upon surface roughness of the conductors and, in the case of coaxial feeders, the dielectric loss. The cross-sections show the relative sizes of the three feeders. It is not practicable to reduce the cable loss at high frequencies by increasing the dimensions because, apart from losing the economic advantages, difficulties due to the possibilities of higher order waveguide modes appear, and greater diameter of cables makes the transitions at the ends more difficult.

Coaxial cables are thus eminently suited to frequencies below about 2 000Mc/s and for systems handling rather small numbers of telephone channels when the end-to-end and internal reflection effects can be tolerated.

Wavguide can be made to meet all the requirements stated previously, and at 3 000Mc/s and above it is the generally used type of feeder. For a given frequency band high order mode effects can be avoided by suitable choice of waveguide dimensions, and well matched transitions to aerials and radio equipment can be made. Its principle disadvantage is its cost and hence it is seldom used for systems carrying small numbers of telephone channels or using low radio frequencies, unless the feeder loss is of paramount importance, such as in a u.h.f. permanent tropospheric forward scatter link. In this case the reflection effects are not of great importance and so cheaper methods of fabrication are possible instead of the more familiar accurately drawn tubing.

Waveguide Feeder Technique

Choice of Dimensions

For a given frequency, the larger the cross-section of waveguide the lower is the loss.

Feeders in general, however, are required to include bends or corners and twist-sections and in these cases if a relatively large rectangular waveguide, or waveguide of elliptical section, were used, unless very great care were taken mode conversion and change of polarization would be produced. Thus unless there are other advantages to be gained, in addition to obtaining a lower feeder loss, it is more practicable to choose a rectangular waveguide in which only the dominant mode of a single polarization will propagate.

Even within this restriction there is a fairly wide range of permissible dimensions if a band of only 400 or 600Mc/s is being considered, and the final choice would be made at the present time from the available series of rectangular waveguides. The intervals in this series are sufficiently close to make two sizes available for any 600Mc/s band, and the smaller size is generally chosen, in spite of the resulting slight increase in attenuation, since it is more convenient for the waveguide components of the radio equipment.

Tolerances

The proposed standard tolerance on the internal dimensions of rectangular guides is one eight-hundredth of the larger dimension.

The reflection coefficient produced at a junction of a waveguide of dimensions $a \times b$ with a waveguide of $(a \pm \Delta a) \times (b \mp \Delta a)$ is given by

$$r = \tfrac{1}{2}\left(a/b + \frac{1}{1 - \lambda^2/4a^2}\right)\Delta a/a \quad \ldots\ldots\ldots \quad (3)$$

$\simeq 2\,\Delta a/a$ for typical values of a, b and λ

$= 0{\cdot}25$ per cent if one guide is on the extremes of both limits in the most harmful way. If each guide is on opposite extreme limits then $r = 0{\cdot}5$ per cent.

The variation of well made waveguide tube within a given batch does not cover the full extent of the tolerances and any variation along a single length is considerably less, and so 0·25 per cent is a reasonable expectation of junction reflections due to aperture errors.

To enable several lengths of waveguide to be connected together in a straight run without deformation there is a limit to the amount of bow permitted in the tube, i.e. the departure from a straight line between any two points on the surface. This is 0·025in edgewise and 0·050in flatwise in a 15ft length.

When two guides are flanged and bolted together there is the possibility of one axis being laterally displaced with respect to the other. Fortunately the reflection due to this type of error adds in quadrature with that due to aperture error, and for the same orders of dimensional error the reflection is smaller by a factor of 5 or 10, but increases roughly as the square of the displacement. By drilling the holes in the flange through a jig which is self-locating on the inside of the waveguide and using either a pair of dowel pins, or bolts with an accurately dimensioned shank, the displacements can be kept less than 0·010in, which for a 2in guide at 4 000Mc/s would result in less than 0·06 per cent reflection.

In avoiding axial displacement, twist at the joint is kept down to a negligible amount, and the remaining possible misalignment, viz angular displacement of the axes, which according to Elson[4] can produce 0·1 per cent to 0·2 per cent per degree, can be held to $\frac{1}{2}°$.

Similarly, the amount of twist in the guide itself is limited to less than 1°/ft at any point, and less than 4° in 15ft, and is taken up by its own flexibility on mounting.

Perhaps more important than the need for accurate alignment, since this can be achieved fairly readily, is the need to avoid damage to the waveguide tube during processing, storage, shipping and installation. It is particularly necessary to take care while the flanges are being hard soldered to the tube that no deformation occurs, either at the aperture or some distance along the tube due to temperature differentials.

Shims

Having ensured the accuracy of the tubing and the alignment of a pair of guides, the remaining requirement is to make good electrical contact along the edges of the aperture. Choke flanges are quite inadequate if a joint reflection of less than 0·5 per cent is the aim, and although it might be obtained by ground contact flanges, it is safer to use a shim of some elastic material, such as phosphor-bronze or beryllium-copper with teeth along the long edges of the aperture. Since s.h.f. radio links usually work with a transmitter power of about 5W, there is no difficulty due to the current density at the teeth. The shim is aligned by dowel pins or precision shank bolts in the flanges, and it can be arranged that under compression the teeth never protrude into the waveguide, nor are more than $d = 0{\cdot}025$in away from the edge of the aperture. The reflection coefficient in a 2in × 0·667in guide at 4 000Mc/s due to this cause for a shim of thickness $t = 0{\cdot}01$in is

$$r = (t/b)\,.\,(2\pi d/\lambda_g) \quad \ldots\ldots\ldots\ldots \quad (4)$$
$$= 0{\cdot}06 \text{ per cent}$$

Since the short edges of the aperture, if not in contact, form non-radiating slots there is no need to provide teeth in the shim along these edges.

Testing

When a piece of waveguide tube has been flanged an electrical check rather than mechanical inspection is preferable, since it can give the reflection performance quickly and directly, and over the frequency band of interest.

A suitable test procedure is to feed the guide by a s.h.f. oscillator which is swept across the frequency band cyclically at a convenient rate. The far end of the guide is closed by a precision termination. Constant fractions of the input wave and reflected wave are mixed in a silicon crystal and the low frequency output is amplified and displayed on a c.r.o. whose horizontal sweep is synchro-

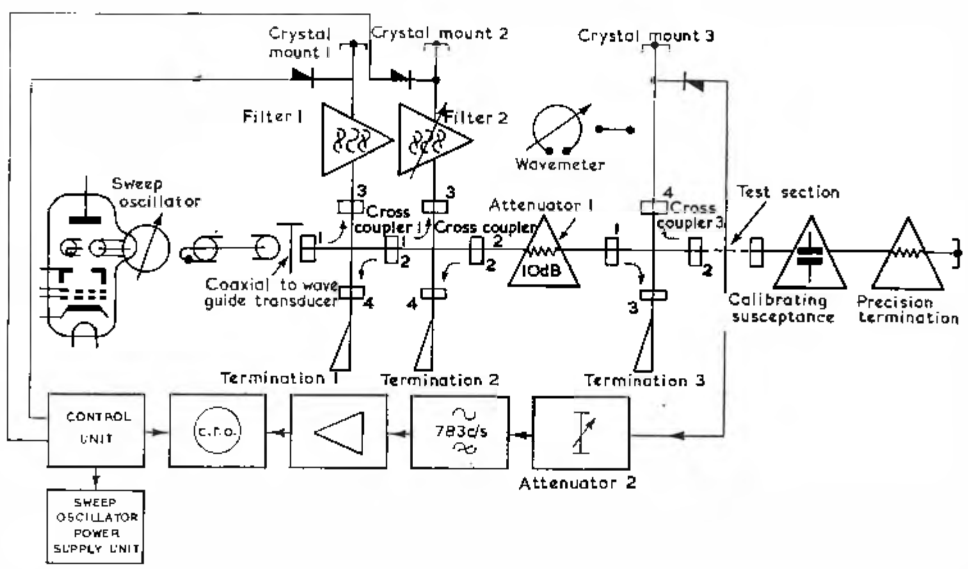


Fig. 2. Microwave wide band frequency sweep reflection measuring set

nized with the sweep of the s.h.f. oscillator. Fig. 2 shows the schematic of such a test set.

Consider a reflection at a distance l from the crystal mixer the oscillator being at a frequency f corresponding to a guide wavelength λ_g.

The phase difference between the reflected and incident waves is

$$\theta = 4\pi l/\lambda_g \quad \ldots\ldots\ldots\ldots\ldots\ (5)$$

As f changes, assuming the phase change at the reflection remains constant,

$$(d\theta/df) = 2\pi \cdot 2l(d/df)\,(1/\lambda_g)$$
$$= 2\pi\,(2l/v_g)$$

Hence, there are $2l/v_g$ cycles of low frequency display per cycle per second frequency shift of the s.h.f. oscillator, and if $v_g = 200\text{m}\mu\text{sec}$ and $l = 5\text{m}$ there is 1/20 cycle of display per megacycle per second shift, i.e. 20Mc/s shift is required per cycle of display.

The oscillator sweeps over a range of 600Mc/s, and therefore a reflection at 5m gives 30 cycles of display and a reflection at 1m gives 6 cycles.

If the phase change at the point of reflection varies with frequency the display is modified accordingly.

Since the reflected wave is of smaller amplitude than the incident wave at the crystal and the incident wave is of constant amplitude, the amplitude of the display is proportional to the amplitude of the reflection.

The set is readily calibrated by placing a screw adjusted to have say 10 per cent reflection in front of the precision termination and reducing the gain of the amplifier before the c.r.o. by 26dB. The envelope of the display obtained then corresponds to 0·5 per cent reflection when the gain is restored to normal. The envelope contour is marked on the screen of the c.r.o.

Faults at the near end of the guide are not displayed very accurately since their corresponding display contains very few cycles, and so each guide is reversed and the ends examined separately.

The usual display is one exhibiting a pure sine wave corresponding to a single reflection at the far end, showing that discontinuities along the guide are relatively very small. The limiting accuracy of the display is determined by the terminating load.

The termination is made by adjusting a pair of nichrome-coated glass vanes in a precisely ground waveguide section. A jig allows the vanes to be moved along the section while maintaining their relative position which is adjusted to produce minimum change in reflection coefficient with a simultaneously low steady value of reflection corresponding to the joint of the section to the apparatus. After adjustment the vanes are locked in position and the termination checked against a slotted-section with a backed-off meter. It is difficult to obtain better than 0·3 per cent across the band of 3 600Mc/s to 4 200Mc/s.

Other components of the feeder, such as corners, twist-sections and gas barriers can be checked by interposing them between a standard 10ft length of waveguide and the precision termination. Since the best possible performance is required from these components, and since it is difficult to consistently achieve better than 1·0 per cent from 3 600 to 4 200Mc/s for a corner with reasonable mechanical tolerances, it is convenient to include three matching screws which can be readily set to obtain the optimum performance (<0·75 per cent) in almost the same time as would be taken to measure the performance of a normal corner.

Installation

The waveguides are supported on a framework connecting the equipment building to the tower, and on the tower itself by means of clamps about 4ft apart, which prevent any rattling of the waveguide in the clamp in a strong wind, and yet do not deform the tubing. The weight of the vertical run is taken on the flange at the bottom of the run, and the layout of the waveguide from the top of the vertical run to the aerial is so arranged that thermal changes in length produce only acceptable amounts of bow in a horizontal member.

The inside surface of the waveguide is protected by a suitable varnish and the flanges are grooved to receive rubber or neoprene gaskets which form a seal between the flange and the shim.

When an installation is completed, air is flushed through the system to remove any condensation, it is flushed again with dry air, and afterwards maintained at a small excess pressure with a drying agent included between a blower and the waveguide system. When there is more than one feeder they are cross-connected at the top so that the humidity of the issuing air can be conveniently checked in the station.

To ensure that no damage has been done during installation, and that there are no other faults leading to poor reflection performance, such as incorrectly fitted gaskets, the feeder is examined by the swept-frequency reflection measuring set in the same way as the individual waveguides were tested. In order to examine long feeders an expanded display is necessary, which allows accurate measurement over a small band which can be located in successive steps across the frequency range of the feeder. The display of a satisfactory feeder is a random superposition of a number of sine waves. A fault shows up as a systematic frequency on the display and by measuring the cycles of display per megacycle per second of s.h.f. oscillator sweep the position of the fault along the feeder can be found.

Modern Trends

Radio links at present installed have a telephone capacity of 600-channels or less, but links are under consideration with a base-band of 8Mc/s to carry 1 800 channels. In order to maintain the performance for the higher base-band frequency, the thermal noise and distortion requirements of the radio system must be 10dB more severe.

One way of reducing the thermal noise is to use feeders of considerably lower loss than hitherto, and to achieve this the waveguide cross-section must be increased beyond the dimensions which guarantee that only one mode is propagated. With such large waveguide care must be taken to avoid the generation of these unwanted higher modes and to preserve the polarization. This precludes the use of corners and twist sections and, in fact, the large cross-section portion of the feeder can only be used in a straight run with carefully designed transitions at the ends.

Since the large guide is capable of transmitting waves mutually cross polarized, it is worth while deliberately using it in this way. In order to obtain small coupling between the two polarizations, it is necessary that for a given frequency the wavelength in the feeder shall be the same for any polarization, which means that the feeder must be of square or circular section. Since it is easier to make tubes of circular section more free from distortion than square section tubes, circular waveguide is used.

To illustrate the magnitudes of the tolerances involved, consider a waveguide of slight ellipticity with a wave polarized at 45° to the axes of the ellipse. At a distance

l the fraction of the field normal to the original plane of polarization is

$$E_L/E = \left(\frac{1 - \cos\theta}{1 + \cos\theta}\right)^{\frac{1}{2}} \simeq \theta/2 = \pi l\,(1/\lambda_{ga} - 1/\lambda_{gb})$$

$$= (l/0{\cdot}543D) \cdot \frac{1}{(n^2 - 1)^{\frac{1}{2}}} \cdot \Delta\lambda_c/\lambda_c$$

where $n = \lambda_c/\lambda$

and D = guide diameter.

If the semi-axes or the ellipse are a and b, $(a-b)/a = 10^{-3}$

$$\lambda_c/\Delta\lambda_c = 1{\cdot}81 \times 10^3$$

then for $(E_L/E) = 10^{-2}$ corresponding to 40dB coupling

$$l \simeq 17\ D^2/\lambda$$

Thus very close tolerances are required to obtain 30dB coupling for the whole feeder, even if one allows random addition of the depolarization effects of the successive lengths of guide.

Another requirement in order to obtain sufficiently low coupling is to launch the waves into the guide with the plane of polarization accurately aligned relative to the bipolar aerial or the polarization separator. If θ is the angular error, the coupling is proportional to $\sin\theta$ and so an error of $1/31{\cdot}6$ rad $= 1{\cdot}8°$ produces 30dB coupling.

When the feeder has been made with sufficient accuracy to avoid the generation of unwanted high order modes and to permit bipolar working, it is feasible to use it simultaneously with more than one frequency band, for example 3 800 to 4 200Mc/s, 6 000 to 6 400Mc/s and perhaps 10 000 to 11 000Mc/s. The difficulty lies not so much in the feeder as in the associated branching circuits.

It is not appropriate to describe here the various techniques that have been suggested for the frequency-band selecting polarization separators, but research work is proceeding to solve the formidable task of preserving the required degree of impedance match over the total frequency range.

The design problems of these separators are still being investigated, and no doubt eventually solutions will be found. It is, however, rather early to make an assessment of the economics of a bipolar multiple frequency band aerial system and since there may also be some reluctance to commit such a large volume of traffic to a single aerial system, it is not certain that this trend will be followed to any appreciable extent.

Acknowledgments

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Colour Television Receiver Development

Research engineers at The General Electric Company's Wembley Laboratories have succeeded in reducing the size of their latest experimental colour television receiver to that of a domestic 21in monochrome set. This represents a scaling-down of nearly 50 per cent on the first model, built three years ago.

Although the research team do not claim that all colour problems have been solved, they have overcome many of the difficulties encountered in combining small size with good performance.

Performance problems were mainly those of definition, registration and fringe signal conditions, consistent with long-term stability. Care had to be taken to ensure that definition was of high quality monochrome standards, and this meant that the colour de-coding circuits had also to be very carefully designed.

The experimental colour receiver (Type TT4) is fitted with an R.C.A. 21in shadow mask tube type 21AXP22A. It has a complement of 35 valves and a total power consumption of 450W. By using metal rectifiers and 'series run' techniques it has been possible to accommodate the receiver in a cabinet measuring 21¼in by 28½in by 31½in.

The receiver power pack incorporates direct rectification of the mains supply to provide one h.t. line of 250V at 200mA, and a second h.t. line of 450V at 400mA is obtained by a voltage doubling circuit. A stabilized negative line of −150V is also available for bias supplies. G.E.C. metal rectifiers are used, and the whole power pack unit measures only 5½in by 7in by 11in. Most of the valves in the receiver are series run, but two small transformers supply the higher current valve heaters and line time-base valves.

The tube e.h.t. supply of 23kV at 1mA is obtained directly from the line flyback without voltage doubling, and a triode shunt regulator is used to stabilize the e.h.t. and thereby minimize the effect of e.h.t. changes on the convergence of the three electron beams.

Passive convergence circuits are fed from the line and frame time-bases and supply the appropriate current waveforms to the dynamic convergence coils mounted round the neck of the tube. D.C. currents are also applied to these coils for static convergence adjustments and all the pre-set convergence controls are easily accessible at the front of the receiver. Line and frame raster shift is controlled by d.c. injection into the scanning coils.

On the signal side, a standard r.f. switchable tuner feeds a slightly modified production type i.f. deck. Sound rejection and sound i.f. take-off conform to usual monochrome practice, but in the vision circuits two separate crystal detectors (GEX54's) are used to provide isolation between the luminance and chrominance channels. This arrangement enables a high definition luminance signal to be maintained, and when this is fed to the ferrite loaded delay line (to provide time coincidence of luminance and chrominance) small reflections in the line are prevented from disturbing the smooth and symmetrical chrominance channel frequency response.

The luminance signal is amplified by two video stages in a 'bootstrap' circuit and is ultimately fed in the correct drive ratios to the three cathodes of the tube. Master brightness is controlled by a bias arrangement in one of the video stages, and the d.c. component of the signal is maintained to within about 1dB.

Two chrominance amplifiers with a 6dB bandwidth of ±500kc/s supply two 'clamping' triodes for high level demodulation along the red and green difference axes, R-Y and G-Y. After filtering, these difference signals are fed to the red and green grids of the tube, and also to a matrix amplifier which forms the B-Y signal for the blue grid of the tube. A reference frequency amplifier provides the two appropriately phased reference signals for the triode demodulators.

A front panel chrominance gain control is provided for adjustment of the colour saturation.

The continuous reference signal required for synchronous demodulation is obtained from an *LC* oscillator which is frequency and phase locked by a two-mode a.p.c. loop, or d.c. quadri-correlator, so that both fast pull-in and good noise performance are achieved.

One of the two detectors in the a.p.c. loop provides a d.c. voltage only when a burst signal is present and when the oscillator is phase locked. Since this is a synchronous detector it provides an accurate indication of the presence of a burst even under adverse signal-to-noise conditions, and is used as a colour 'killer' control for switching off the chrominance channel automatically when a burst signal is absent. The output of this detector is also fed as an automatic chrominance control or a.c.c. signal to bias the chrominance amplifier from which the burst output is taken.

The colour burst is separated from the chrominance waveform and amplified before being applied to the a.p.c. detectors. The separation is carried out by a gating circuit which switches on the burst amplifier only during the burst period of about 4μsec. In order to be quite independent of line time-base synchronization, the gating circuit operates from the back edge of the line sync pulses appearing in the sync separator output.

Percussion Circuits for Electronic Musical Instruments

By Alan Douglas, M.I.R.E., M.Amer.I.E.E.

All keyboard electrical tone producers are essentially steady state sources. After the key or other initiating control is operated, no real modification of this state is possible. A valuable form of envelope control is provided by percussion circuits.

IN spite of changes in harmonic content, loudness level and vibrato brought about by the skill of the performer, all instruments not strictly of the organ type tend to sound uninteresting after a short period of time. This applies particularly to the melodic devices employing miniature keyboards.

If a little thought is given to which characteristic of a musical sound really determines its appeal, it will be found that the shape of the envelope containing the waveform makes by far the most important contribution to the whole effect. In organ-like instruments it is not necessary to take this greatly into account, because if the start of the sound is delayed for a few milliseconds then it agrees very well with pipes; and although the sound stops more rapidly than that from pipes (for a resonator does not release its energy immediately), yet the discrepancy is usually masked by reverberation so that the envelope control problem is more readily solved. The pipe must have a fixed envelope because the player can only admit air under a fixed pressure, he cannot vary the energy supplied to the resonating tube.

But in orchestral instruments we find a very different case. The range of envelope control is dependent on the degree of elasticity and density of the vibrating element and the amount of power which has to be transferred from this to the resonator to excite it. Thus in instruments using an air stream we find extreme elasticity, resulting in a comparatively slow rate of attack or tonal initiation unless the pressure is very high; in stringed instruments the density of the system is greater and so it can be set into vibration more rapidly; and in an extreme case such as the xylophone or triangle, the rate of exchange of energy is very great and the sound is instantaneous. This process of setting the resonator in motion is quite separate from the state obtaining when in full vibration, although it is closely connected with the production of tones.

It is clear that when a performer has complete control over the rate and degree of energy transfer, he can produce a wide range of tonal states but, in general, he cannot alter the overall characteristic tone quality of the coupled system; what he can change very greatly (in many cases) is the rate of exchange of the stimulating force, in other words, the attack.

Deprived of attack, it is often very difficult to identify the kind of instrument producing the sound, and although the rate of decay is also contributory to the envelope it is the initiation of the tone which is important.

Owing to the fact that in an electrical system the latent waveform already contains all the harmonics required to form the complete tone-colour, control over the attack does not produce the same effect as, say, that of plucking a string; since in the latter case the system is at rest and is selectively excited by the act of plucking; but it is doubtful whether the ear can follow the building up of the tone with a plucked string in the upper registers; and for the lower notes, the resonator is often not fully energized so that the whole tonal spectrum is not fully developed. Faced with these conditions, simulation by electrical percussion circuits is feasible; although if we take the case of the plucked string again, as the force applied to the string is increased the harmonics produced also increase together with the loudness; whereas in an electrical method only the loudness will increase. Therefore electrical percussion is most effective with simple waveforms.

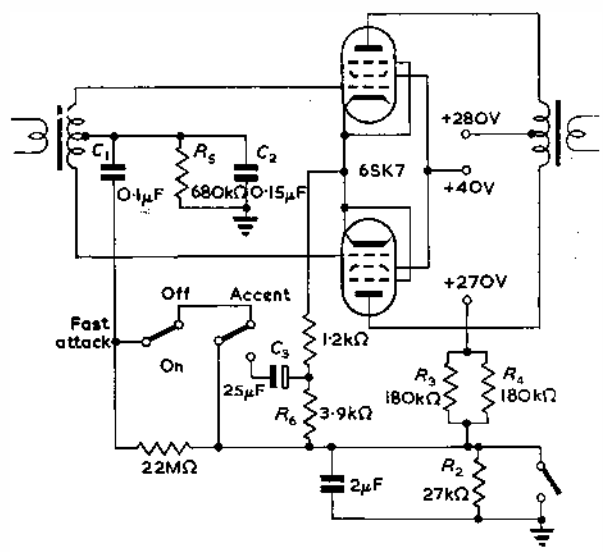


Fig. 1. Push-pull percussion circuit

For obvious reasons it is more economic to apply percussion to melodic instruments. In the Novachord it could be applied to chords as well. Many circuits have been used or proposed for this purpose, all based on the holding of a valve in a cut-off condition with means for causing it to conduct in a particular way when required. In some cases the signal can also be made to decay under control, in others a steady state is reached and held. Yet other circuits merely control the decay rate or even cause the note to be sustained for a time—a form of long decay. These latter circuits are generally used to simulate reverberation in a crude way.

Taking one of the early arrangements for providing envelope attack control, we see in Fig. 1 that the signal is fed by means of a transformer to a push-pull amplifier. This is desirable in such a circuit to reduce clicks or thumps since the push-pull drive removes these. The 6SK7 pentodes are normally held cut off because their grids are earthed and the cathodes biased to about 65V positive by the voltage divider R_2, R_3, R_4. The keying circuit, which is common to and operated by every playing key, earths the potential divider and makes the valves conduct. This would result in a very abrupt attack were it not for C_1, C_2 and R_5. The sudden reduction in cathode voltage caused by keying causes a negative surge through C_1, charges C_2 negatively and moves the grid voltage momentarily in a negative direction. This maintains the cut-off condition for an instant after keying, until the charge on C_2 leaks away through R_5 and thus an agreeably slow start is given to the sound.

However, the instantaneous attack can be used by operating the fast attack switch which disconnects C_1. The 'accent' switch ingeniously gives a percussion effect by connecting C_3 across R_6, while removing C_1 from the circuit. When the cathode cut-off voltage is suddenly removed

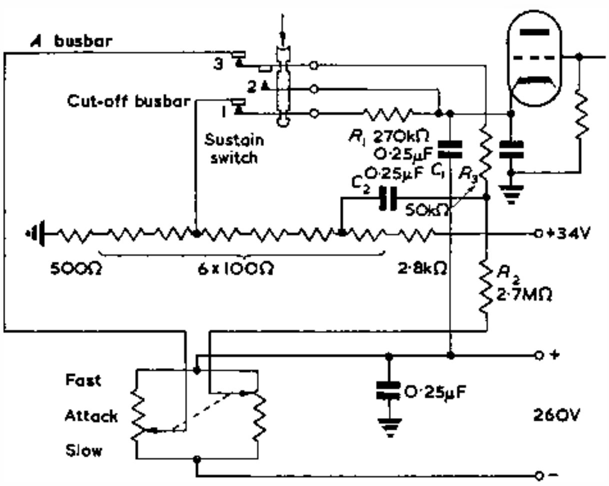


Fig. 2. Infinitely variable percussion circuit

by keying, C_3 effectively short-circuits R_6 and reduces the bias for a moment, so causing the sound to be loud at first and then rapidly die down to the normal sounding level. Of course, with one fixed value of C_3 for all notes, the time-constant cannot be agreeable for the whole register, but in general (and especially in rapid passages) the effect is very good.

Turning to a more flexible arrangement, we see in Fig. 2 that the mechanism for the keys requires three contacts. In addition, isolated supplies of 260V and 34V are required. This latter provides the cathode cut-off bias and some adjustment is available. The switch marked 'attack' varies the potential of A between earth and -260V. The potential to which A is set fixes the initial intensity while the key is held down. Thus, if A is at high potential and C at low potential, the attack will be percussive, the note decaying to a low level. Reversing the potential gradient causes the note to start low and build up. If both busbars are at the same potential, the note will have constant amplitude and be sustained.

The 6W7G control valve is cut off by the voltage from the 34V source through R_1. On pressing a key, contact 3 leaves A and touches 2, while 1 breaks contact with the cut-off busbar. The voltage applied to contact 2 charges C_1 and is transferred to the cathode of the valve, which conducts. On releasing the key, the note decays quickly as C_1 discharges through R_1 unless the sustaining switch is operated. As can be seen, this prevents contact 1 from closing, which means that the decay is slow as C_1 can then only discharge through the control valve.

The initial potential of C_2 is fixed principally by the setting of A, as R_3 is smaller than R_2. With the attack switch at 'fast', A will be at high potential and on keying, the note will sound suddenly until the potential of C_2 approaches that of C. As indicated above, if the conditions are reversed by setting the attack switch at 'slow', the note

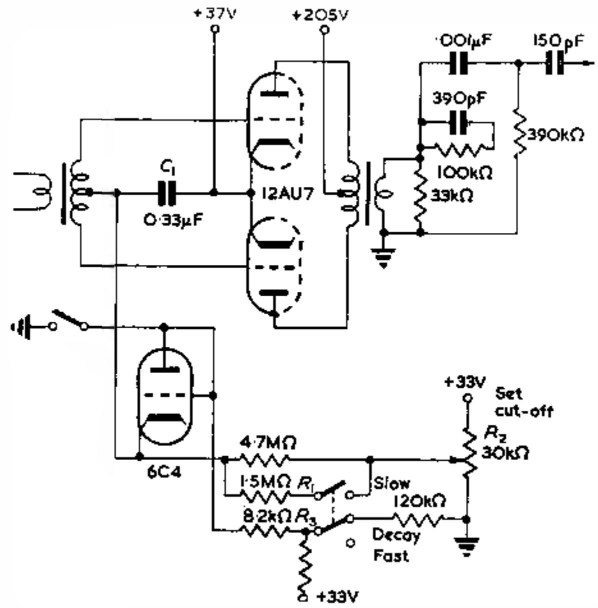


Fig. 3. Decay circuit

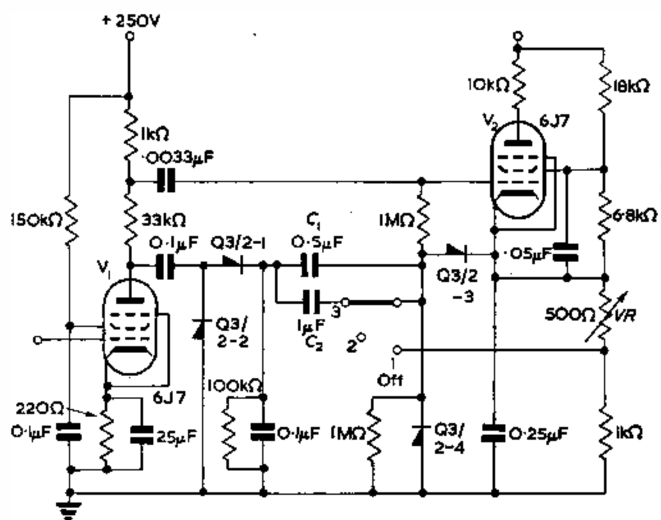


Fig. 4. Self-triggering percussion circuit

starts at a low intensity and builds up as C_2 charges. Although somewhat complicated, the control offered by this circuit is very wide and some remarkable effects can be obtained.

In the case of a tone quality which has no very high harmonics, the start of which may be made tolerable by merely inserting a simple RC network to reduce the click, a fairly steep wavefront is to be expected. It may then only be necessary to shape the decay of the tone. Such a circuit is shown in Fig. 3. A push-pull amplifier is fed through a transformer. The cathodes are cut off by a positive bias which can be neutralized when the 6C4 (connected as a diode) conducts. The level can be set by R_2. On keying the diode (which may be performed by a waveform if of sufficient amplitude) the 12AU7's conduct and the rate of decay is set by R_1, R_3 and C_1. A filter removes any trace of 'thump' due to the diode.

Obviously, if a separate keying contact could be abolished, the potentialities of this kind of circuit could be increased. One proposed arrangement is shown in Fig. 4. The object here is for the incoming signal to generate a d.c. charging pulse which is superimposed on the tone so as to render it percussive. The upper line from the anode of V_1 carries the normal signal and if the switch is at 'off', amplification takes place in V_2 in the ordinary way, the two 1MΩ resistors acting as the grid resistor. *VR* is set up to provide the proper level of conduction. With percussion 'on', a d.c. charging voltage is provided at Q3/2–1 and the 100kΩ resistor and 0·1μF capacitor produce a mean d.c. level from the fluctuating signal. Whichever capacitor is in circuit discharges into the cathode of V_2, causing a sudden rise in voltage which gives the percussive attack. At position 2, the 1MΩ resistor discharges C_1 fairly rapidly, but in position 3, the charge can only leak

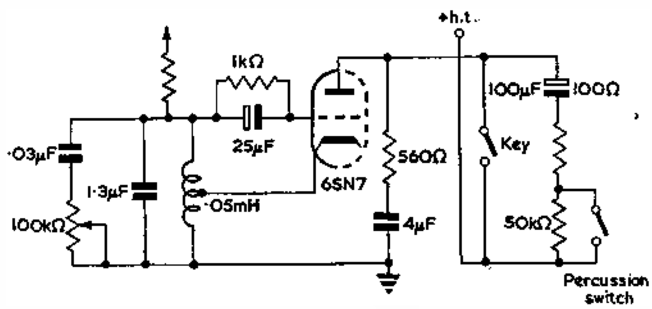


Fig. 5. Oscillator for chime effect

away slowly through Q3/2–4. As the charge on C_1 or C_2 is practically a square pulse, Q3/2–3 and Q3/2–4 assist in removing excessive transients, since after a certain level they conduct and are so poled that the charging and discharging transients are reduced.

An interesting arrangement is one where the self-inductance of the tone oscillator is very great, due to the use of an air-core coil. In such a case, the oscillator is slow in starting and slow in stopping, and this effect can be increased by a large grid capacitor. If, then, an additional time-constant is introduced into the +h.t. keying circuit, as in Fig. 5, a quite harp-like effect can be produced. This is not a non-linear signal as in the circuits previously described, but both the rise and decay times are similar. For chime and similar tone envelopes it is effective.

Reverting to the simplest case, which is not entirely free from clicks, either a valve of the 6BA6 type or a small thyratron can be used as in Fig. 6. The fixed cut-off bias is removed on earthing the positive line but the *RC* networks shown produce a reasonable percussive effect.

An alternative method suitable for the amateur and which does not impose any appreciable drain on the bias supply is shown in Fig. 7. Here, the screen grid of the valve is negatively biased so that no signal passes. On closing the common keying contact shown as a selector switch, the potential of the screen is raised to that of the +h.t. line and the valve conducts. By choosing suitable values for R_1, R_2 and R_3 and C_1, a variety of attack states is possible and if R_1 is quite low, a percussive effect obtains.

All the foregoing circuits depend on the ratios of *R* and *C* in charging networks and all are adjustable over a range. Preferred values shown are in some cases found to be those most suitable for a particular application of the instrument in which they are used, or for some specific waveform fed thereto.

In order to reduce the complexity of keying contacts, recent developments have resulted in making use of the d.c. pulse created when an oscillator is used, an illustration of which is shown in Fig. 8. Similar circuits are used by Baldwin and Dorf in the Schober organ.

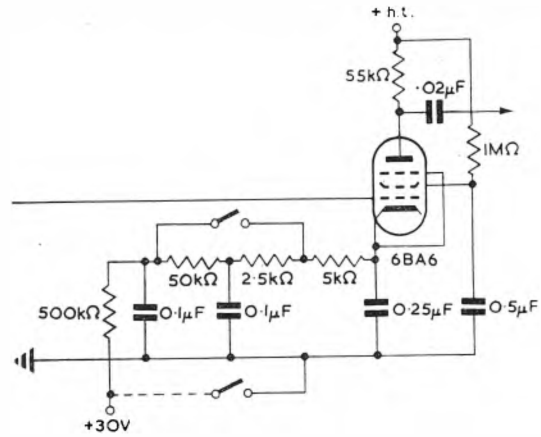


Fig. 6. Elementary percussion circuit

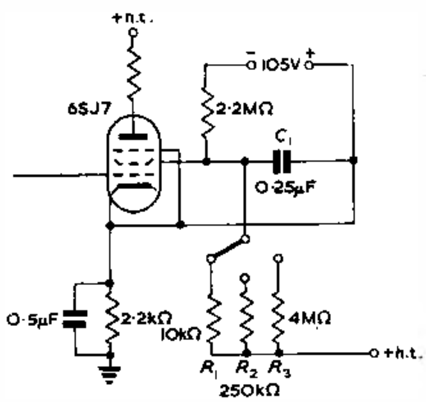


Fig. 7. Experimental percussion circuit

The h.t. supply to the two music generators shown at the bottom left hand of Fig. 8 derive their supply through resistor *A*. On keying, a voltage drop occurs across *A* and the pulse resulting from this is transmitted through *B* and *C* to the amplifier which is really a cathode-coupled multivibrator with time-constants fixed by the values of *E* and *F*. This is so that the relay *D* will have time to operate properly. So soon as the voltage decays through *E* and *F*, the amplifier returns to a cut-off state and the relay drops out; therefore the contacts shown coupled by the dotted line close only for a moment. Now if the upper amplifier *H* is cut off by a bias voltage applied at –*CO*, no signal can appear at its output even if the generators are switched to the input. But when the relay closes, the contact earths the valve grids through the resistors shown, which allows the valves to conduct rapidly. The sound is heard abruptly and would continue at a steady level. However, the relay opens almost at once, so that cut-off bias is again applied through the other resistor and the

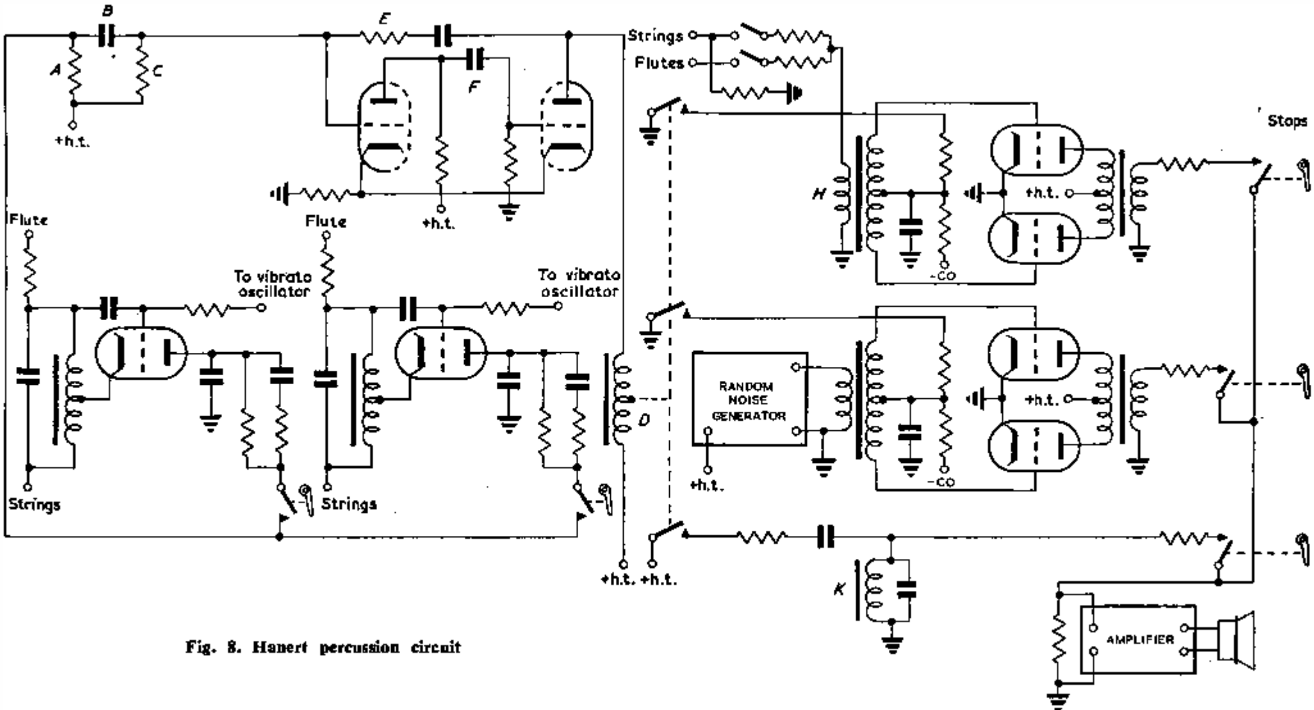


Fig. 8. Hanert percussion circuit

associated capacitor; the time-constants of this circuit allow the signal to die away more slowly, so that the effect is something like that from a plucked string.

At the same time, a continuously-operating noise generator is applied to another amplifier with similar time constants, so that a pulse of noise is also injected. This gives a greater sense of realism to flute-like tones.

An ingenious idea is seen at *K*. The resonant ringing circuit shown is energized through a capacitor with a pulse from the +h.t. line; this causes a thump and when added to a musical signal has something of the nature of a 'chinese block' attack. Any or all of these effects can be switched in at will by the 'stops' shown at the extreme right side of the circuit.

In all the foregoing circuits it must be borne in mind that many orchestral sound are not, and cannot be made, percussive; therefore one must not expect satisfactory results on this kind of waveform.

Production of Ferro-Magnetic Memory Cores

The Chemical and Metallurgical Division of the Plessey Co. Ltd has recently moved into a new factory at Towcester with increased facilities for the production of square loop ferrite cores for use in digital computers.

The first stage in the production consists in the mixing and granulating of the basic components (iron, manganese, copper etc.) after which the mix is introduced into moulds and pressed to the correct density, size and weight.

The resultant compacts, which can then be handled, are fed slowly through a furnace, the temperature of which is accurately maintained. The finished magnetic cores are then removed and tested.

At Towcester all cores are double-tested as this is an extremely important step in the memory fabrication process. A check is made on temperature characteristics of cores as well as their output voltages and switch time characteristics.

It will be apparent that a considerable amount of effort is involved in the twofold inspection of individual cores when consideration is given to the fact that a single 64 × 64 matrix uses 4096 cores. The calibration of the testing unit must be extremely accurate, as core testing is a continuous operation.

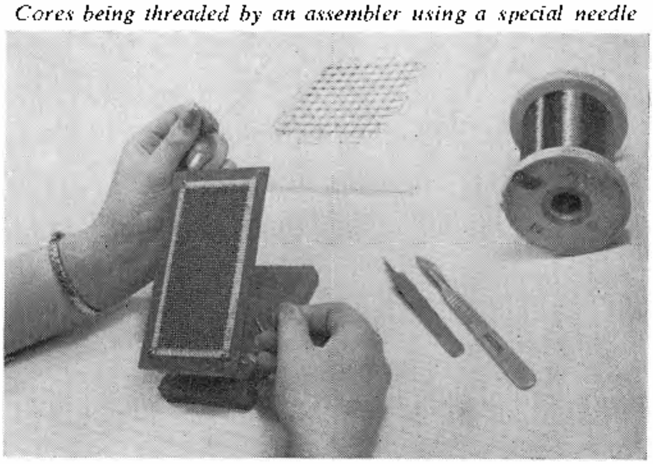
After testing, the cores are assembled on jigs, and the vertical and horizontal wires are fed through the ultra-small cores by hand. This job demands a keen eye and considerable dexterity on the part of the assembler, as the smallest core is only 0·050in o.d. and 0·030in i.d. These small cores have made possible the construction of matrices for operation with transistors, while still maintaining their high speeds.

There are two distinct grades of ferrites produced by Plessey for use in magnetic memories; these are "Ferramic", being the trade name for an improved grade of ferrite material.

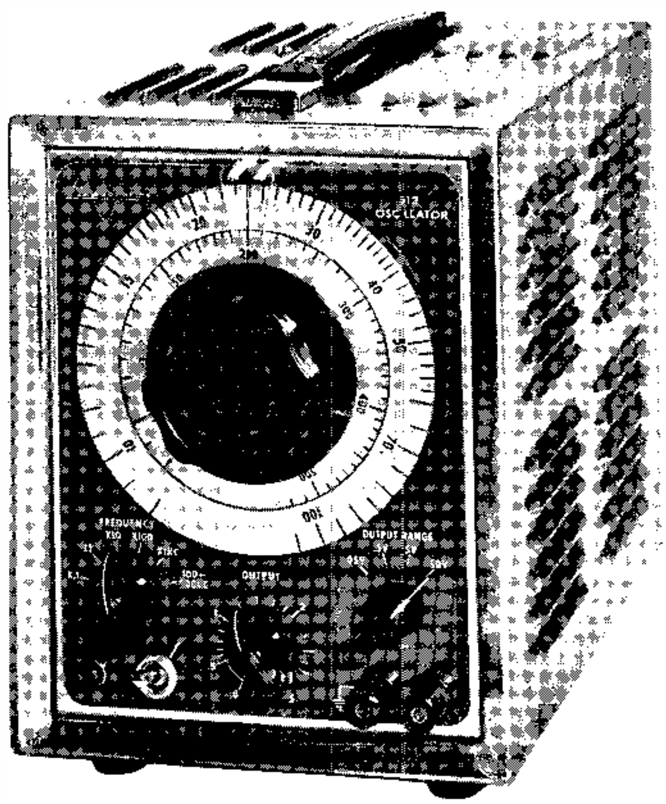
The particular features for which Ferramic S4 is notable are high operating speed, high output voltages on switching, and low values of noise voltage outputs, whereas the S5 material is suitable for applications where a lower operating current is required. Apart from their widespread use in memory cores for high speed, random access, computer memories, the cores can be used for various logical and switching circuits in computers. Multi-aperture cores in square-loop material have also been developed by Plessey, and these can be used in shift registers and gating circuits, as well as for non-destructive readout stores.

Recent investigations at Towcester have been on magnetic and super-conductivity devices, the principal objectives being to effect increased speeds, in the order of ten times and reductions in the operating currents of memory cores.

Cores being threaded by an assembler using a special needle



A Wide Range Sine Wave Generator



By Leon H. Dülberger* and Howard T. Sterling†

A Wien bridge oscillator provides a very low distortion sine wave over frequency range of 0·9c/s to 510kc/s. An output power amplifier employing single-ended push-pull circuits is provided. Harmonic distortion is under 0·15 per cent when driving a 600Ω load at a 2W level. For loads of 1 200Ω or greater, 50V r.m.s. is available at comparable distortion. Frequency stability is within 0·5 per cent for wide temperature and line voltage variations.

THE problem of generator design to provide signals over a frequency range from 1c/s to 500kc/s often resolves itself in the form of several specialized units, particularly when requirements of high power and low distortion are also imposed. One is accustomed to working with equipment in which compromises in distortion and output power have been made to achieve this wide frequency range. At higher power levels distortion content often rises in direct proportion to power output unless the bulk of the instrument approaches unhandy proportions.

Through the use of refined techniques, integration of all these considerations can be accomplished in one compact unit. It is the purpose of this article to describe a wide range generator of relatively high power for low impedance operation with distortion components low enough to be discounted in most measurements.

The frequency range covered by the instrument extends from 0·9c/s to 510kc/s, in six tuning ranges. Two watts of power are developed into a 600Ω load with total harmonic distortion below 0·1 per cent over most of the operating range. The complete instrument consists of a Wien bridge low distortion oscillator, a single-ended push-pull output amplifier with heavy feedback, and regulated power supply.

Oscillator Section

The popular Wien bridge circuit is used as shown in Fig. 1. Five decade ranges are provided from 0·9 to 11c/s to 9kc/s to 110kc/s. A sixth bandspread range provides improved readability from 90kc/s to 510kc/s. These are normally calibrated to an accuracy of ± 2 per cent of dial indication. Operation is explained as follows. Positive feedback from the anodes of V_2, V_3 is applied to the grid of V_1, through the *RC* arms of the Wien bridge, resulting in oscillation at the bridge controlled frequency.

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† *Waveforms Inc., U.S.A.*

Amplitude of oscillation is determined by the magnitude of negative feedback applied to the cathode of V_1. This is derived from the divider comprising R_1, R_2 and the lamp. Amplitude stability is controlled by the 10W, 220V lamp in the lower end of this divider. Its resistance/temperature characteristic is such as to oppose a rise in signal voltage from V_2, V_3 by effectively increasing the amount of inverse feedback to the cathode of V_1. Selection of the operating point of the amplitude-control lamp must be made with a view to final distortion in the oscillator. This is particularly important when using this configuration at very low frequencies. At 1c/s, recovery time of the lamp will determine stability and waveform purity to a great extent. Optimum oscillation level is found just below the knee of the distortion/amplitude curve at mid-frequency. In this configuration, it is about 20V r.m.s. at the anodes V_2, V_3.

All components in the negative feedback loop must be free of voltage and temperature coefficient error. Ordinary composition resistors at R_1 and R_2 have been known to introduce appreciable distortion.

The oscillator of Fig. 1 normally operates with a mid-frequency distortion of 0·05 per cent, rising toward the frequency extremes.

Other factors which contribute to waveform purity are the choice of valves suited to the design. The input valve type is dictated by limitations of grid conduction. This requirement is imposed by the high values of grid circuit resistance selected to obtain very low frequency coverage. Even trivial current flow in the grid circuit will result in non-linear loading of the tuning circuit. Type 6SJ7 was found to be superior in this respect, and was chosen for this reason.

The output valves, V_1 and V_2 are type 6AK6, selected for their high linear anode signal capability, for low cathode current. They provide relatively high g_m commensurate with low distortion and provide sufficient open loop gain to assure heavy degeneration.

By disconnecting the bridge circuit *RC* arms, the oscillator operates as a two stage amplifier as outlined in Fig. 2, and simplifies optimization, i.e., correct selection of operating points and load impedances. Separate cathode resistors and, as a corollary, bypass capacitors, provide independent degeneration for each 6AK6 to stabilize d.c. operating points. Hence, each tube contributes essentially equal output current. To a great extent, differences both in initial g_m and in aging of V_2 and V_3 will be compensated. Parallel connexion of valve anodes is made through a 100Ω resistor to prevent parasitic oscillations.

Amplitude of the generated signal and frequency against range (90kc/s to 510kc/s) a 0·01μF capacitor, C_3, is switched in from screen to cathode to provide higher loop gain independent of frequency.

The 5:1 tuning range (90kc/s to 510kc/s) is achieved by switching out two of the four sections of the main tuning capacitor and including range-restricting capacitors. The resulting increase in the impedance of the tuning network reduces the high frequency loading on V_2, V_3. Use of unequal resistance in the frequency determining arms of the bridge circuit corrects for a residual drop in amplitude which would otherwise result from loading on this range.

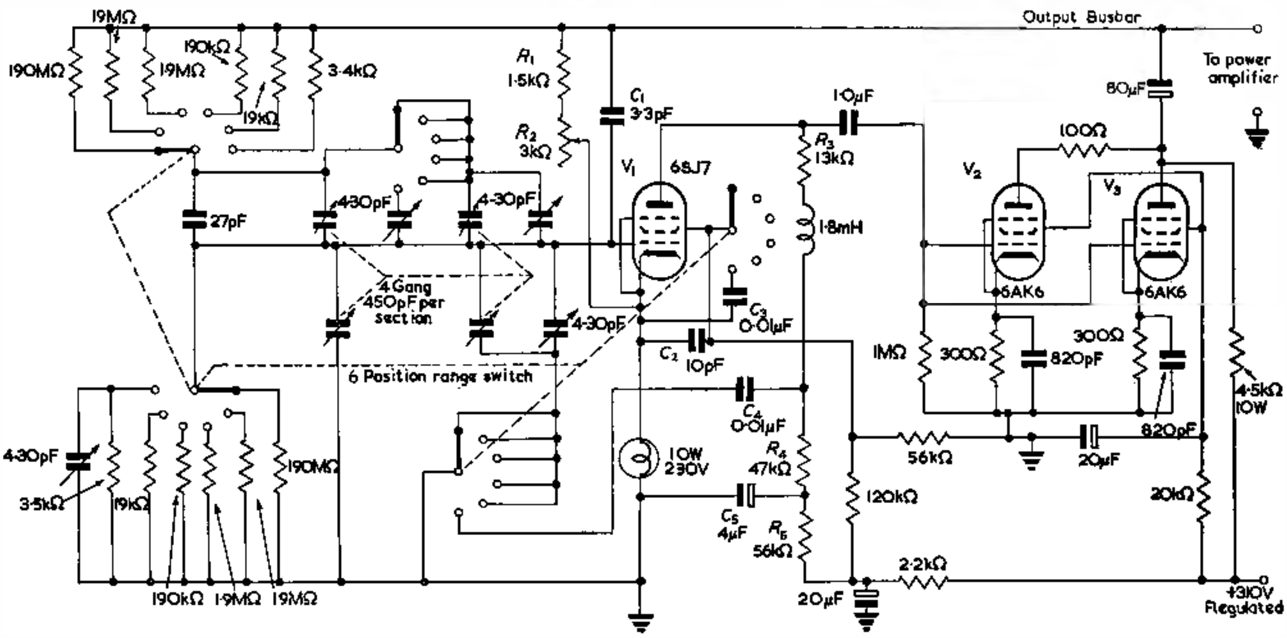


Fig. 1. The wide range generator circuit

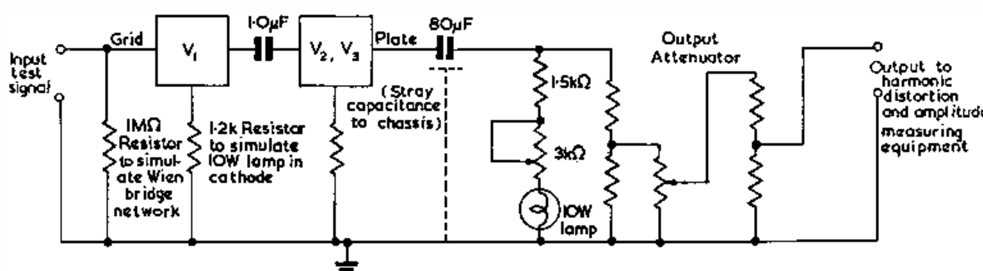


Fig. 2. Operation as two-stage amplifier

dial rotation from range to range both tend to remain constant through 80kc/s. Beyond this point, measures must be taken to prevent errors due to phase shifts in the oscillating amplifier. Stray capacitance in valves and wiring produces a roll-off which results in a lagging phase angle around the loop. To offset this effect, C_1 is connected from the output bus to the control grid of V_1. This introduces a small leading current. Shunt peaking is also introduced in the anode circuit of V_1 and is effective around 150kc/s and above. A rising characteristic with frequency is provided by the selection of cathode bypass capacitors in the output stage. They are chosen to reduce degeneration, beginning at about 100kc/s.

Capacitor C_2 from the screen of V_1 to its cathode reduces screen degeneration at high frequencies and compensates for the effects of stray capacitance elsewhere. On the last range the anode load of the input stage is reduced to R_3 (13kΩ), shunt peaked on this range, by switching in C_4 to effectively bypass the 47kΩ anode load resistor R_4, which forms the major load on other ranges.

At low frequencies a phase shift problem is again encountered due to amplifier roll-off. This produces a slight 'pull' of dial calibration and increased distortion on the lowest range. A compensating circuit comprises C_5 and R_5 in the anode load of V_1. The reactance of this capacitor rises to provide roughly a 3dB 'hump' at 1c/s. This corrects for the low frequency phase shift elsewhere, and provides corrected dial tracking.

Power Amplifier

An amplifier follows the Wien bridge oscillator in order to provide sufficient power into an external low impedance

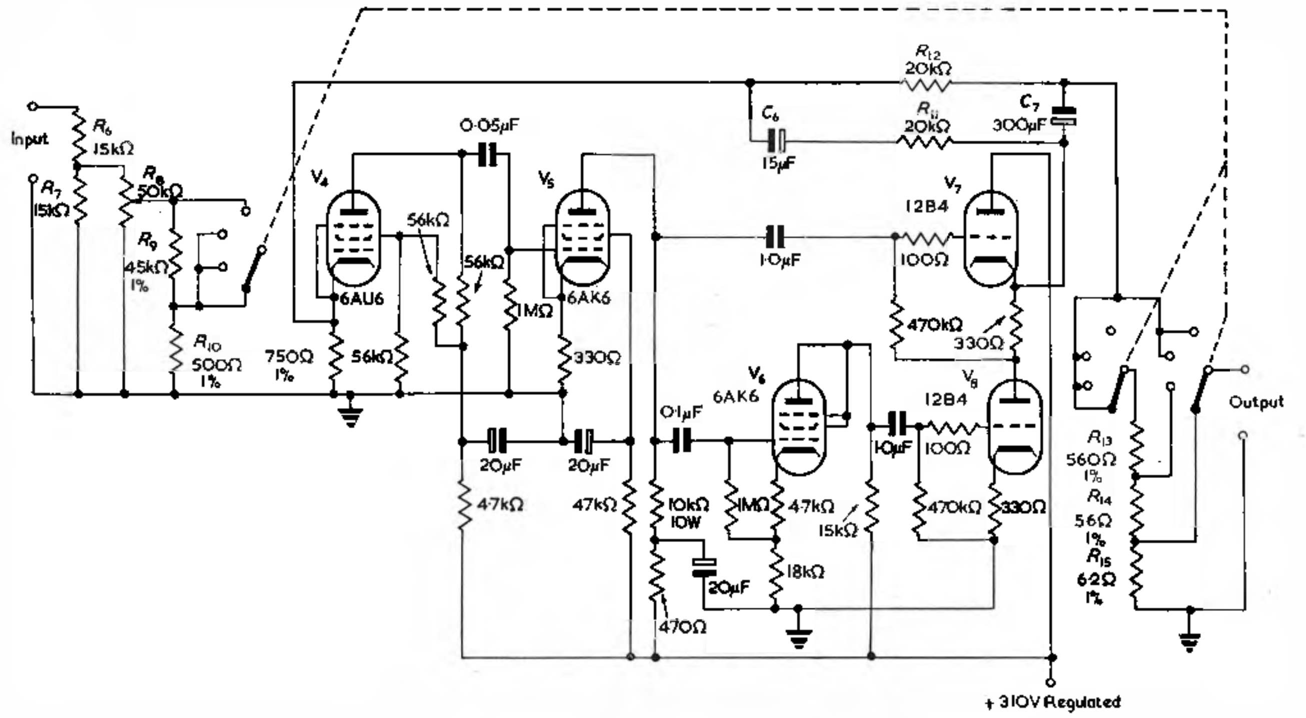


Fig. 3. The amplifier circuit

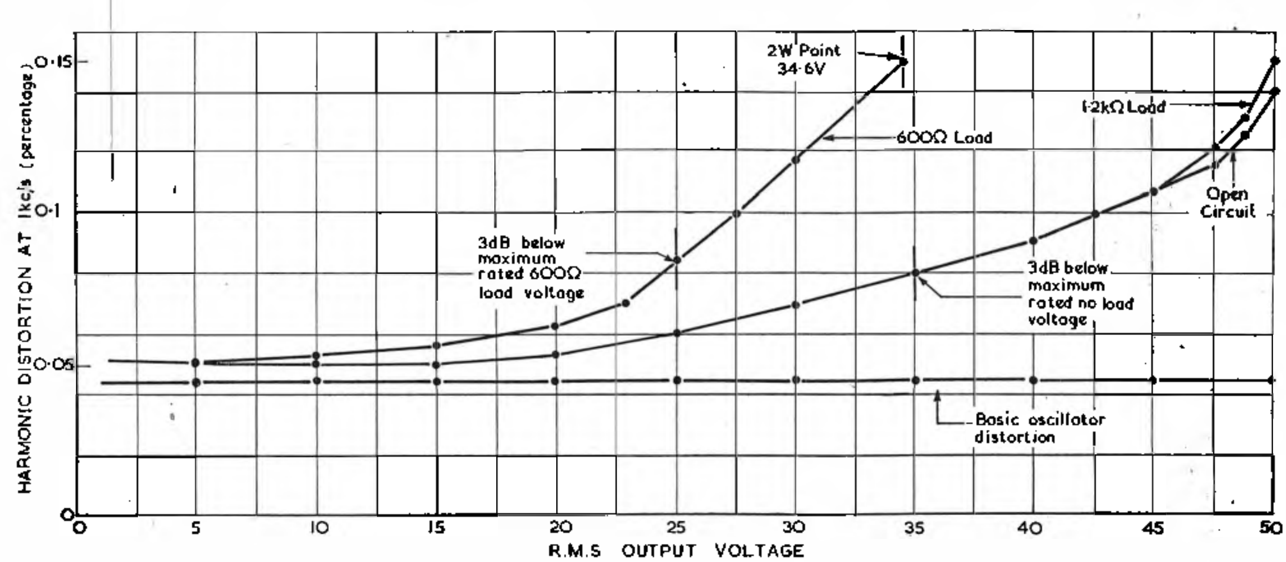
load. The amplifier consists of a voltage amplifier, driver, phase invertor, and a single-ended push-pull output stage. The circuit is shown in Fig. 3.

A full two watts is delivered to a 600Ω load (34·6V r.m.s.). An 'open-circuit' voltage of 50V r.m.s. is available into loads of 1 200Ω or more. Distortion content for various load and signal level conditions is shown in the curves of Fig. 4. It will be noted that the distortion is well under 0·1 per cent for all levels 3dB or more below maximum, and is only slightly greater at rated output. Careful selection of operating parameters makes this performance possible.

Output from the oscillator is applied through a resistive network to a nominally linear potentiometer. For a calibrated 10 : 1 range of output voltage, a linear control produces excessive compression. A logarithmic characteristic was considered desirable to improve adjustment accuracy. To provide the required stability of adjustment and calibration a wire-wound unit is called for. All these requirements are met with a conventional wire-wound control in the following manner. The slider of R_8 is loaded by the input attenuator resistors R_9 and R_{10}. In conjunction with the input network R_6 and R_7 they modify the output of the potentiometer to an essentially logarithmic function of rotation. The curve of Fig. 5 shows the relation of output voltage to percentage resistance of this control.

A 6AU6, V_4, input voltage amplifier, selected for low noise and high gain, feeds a 6AK6 driver stage, V_5. Type 6AK6 was chosen here for the same reasons as in the oscillator (above). The driver output is applied to the cathode-follower V_7 and also to the phase invertor driver

Fig. 4. Distortion content for various load and signal level conditions



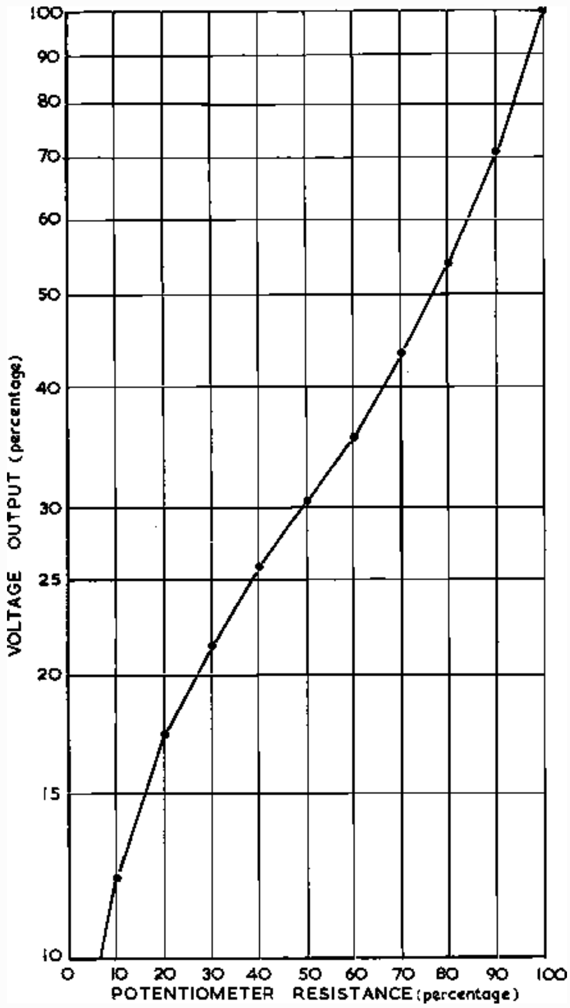


Fig. 5. Output voltage against percentage resistance of amplitude potentiometer

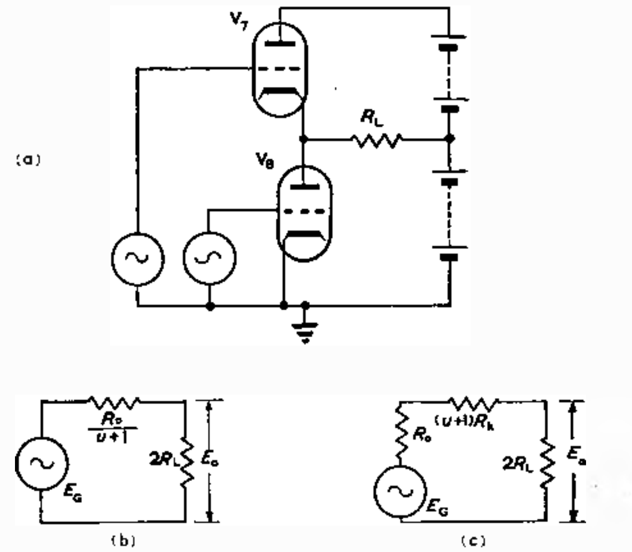


Fig. 6. The output circuit and its equivalent circuits

6AK6, V_6, which provides signal in correct phase and amplitude to drive the lower output valve V_8. Balanced output into the load is thus provided.

The output circuit as detailed in Fig. 6(a) is shown to combine a cathode-follower and an anode loaded amplifier under a.c. signal conditions. The control grids are driven 180° out of phase. Amplifier action can be clearly seen to be true push-pull, with anode current in V_7 rising while that of V_8 falls. D.C. relationships are not shown in Fig. 4.

Gain of the output circuit is of the order of unity and can be roughly calculated from the equivalent circuits of Fig. 6.

For the cathode-follower, V_7, Fig. 6(b):

$$A = (E_o/E_G) = \frac{2R_L}{R_a/(\mu+1) + 2R_L}$$

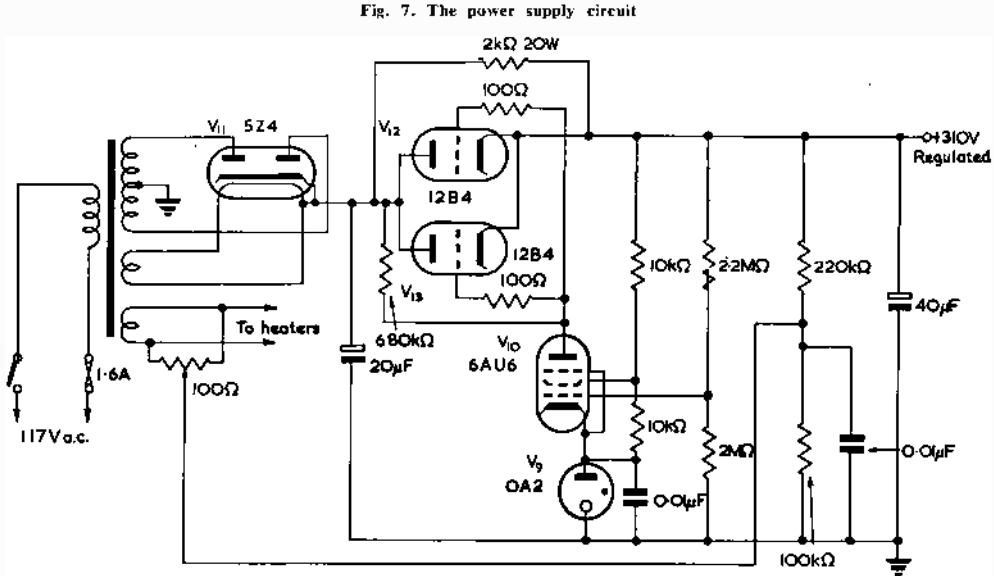
For the anode loaded tube, V_8, Fig. 6(c):

$$A = (E_o/E_G) = \mu \frac{2R_L}{R_a + (\mu+1)R_K + 2R_L}$$

Separate bias resistors are used for each valve to stabilize the operating points of the individual valve.

Negative feedback is taken over the entire amplifier from the output stage to the cathode of V_4. The 300μF output blocking capacitor C_7 is included in a secondary loop to

Fig. 7. The power supply circuit



reduce its effective reactance. To keep d.c. voltage at the output terminals (from the cathode of V_4) to a minimum most of the feedback is taken from the positive side of the 300μF capacitor C_7, through a 20kΩ resistor R_{11} and 15μF capacitor C_6. An additional 6dB of feedback is routed through the 20kΩ resistor R_{12} connected to the output side of the 300μF capacitor. This configuration increases the effective coupling capacitance to an equivalent value measured at 1 050μF.

Four decade ranges of output voltage are provided at the output terminals. Each is continuously variable by means of the calibrated amplitude adjust potentiometer. These ranges are calculated to realize the best signal-to-noise (and hum) ratio in the amplifier for any amplitude setting.

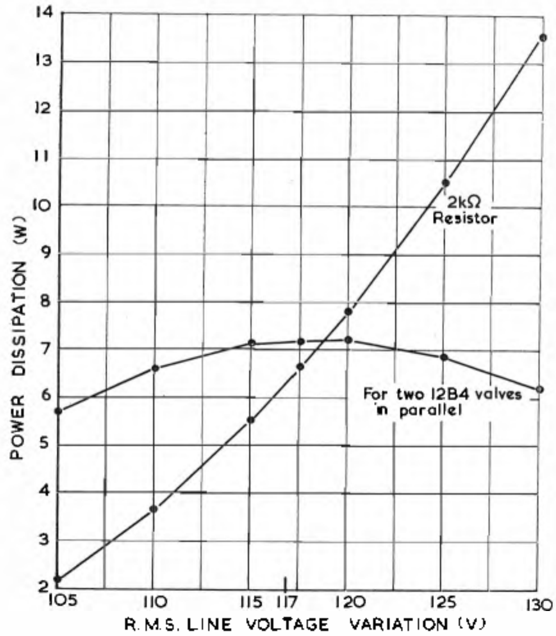


Fig. 8. Division of power for control valves and resistor

Signal into the amplifier is unattenuated at both the input and output on the 50V range. For the 5V range the input divider is switched down 20dB in voltage and the series connected output attenuator resistors switched across the output. They serve to shunt any d.c. leakage through C_7 on this range. For the 0·5 and 0·05V output ranges the input level is held at this same value, while two 20dB steps are selected at the output attenuator. The junctions of R_{13} and R_{14}, and then of R_{14} and R_{15} are selected respectively.

Dynamic output impedance of the amplifier proper is equivalent to 6Ω in series with 1 000μF on the 50 and 5V ranges. Source impedance on the 0·5V range is approximately 60Ω, and 6Ω on the 0·05V range. This low impedance provides essentially constant voltage output with most external loading and insures operational stability. Stability is maintained in the presence of very low resistance (virtual short-circuit) and reactive loads.

Power Supply

In equipment operating at such a low frequency, variations in power supply voltage may be amplified in signal circuits. Accordingly, it was found necessary to provide an electronically regulated power supply, detailed in Fig. 7. It should be noted that the regulator must be capable of accommodating the full swing of signal current of the output and oscillator amplifiers at low frequency. Two 12B4 tubes in parallel serve as the series control element. A 2kΩ-20 watt resistor in parallel with these tubes assures operation within anode dissipation limits. Fig. 8 shows dissipation for both valves and resistor over the rated range of line voltage.

Extensive decoupling is provided to the low level circuits to prevent instability due to common impedances.

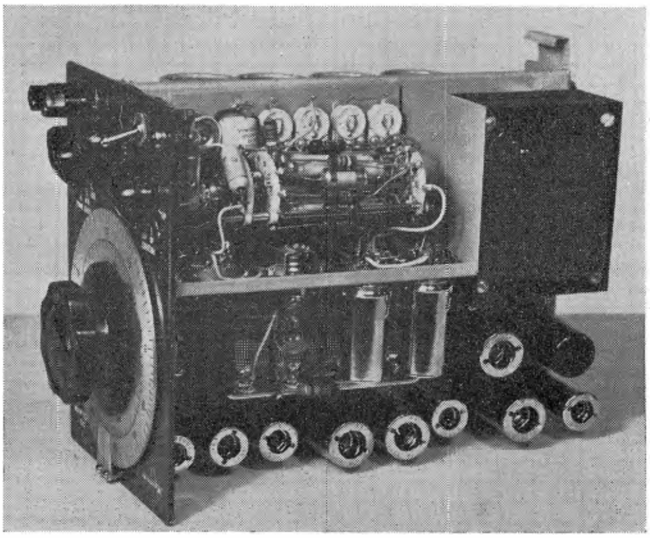


Fig. 9. An internal view of the generator

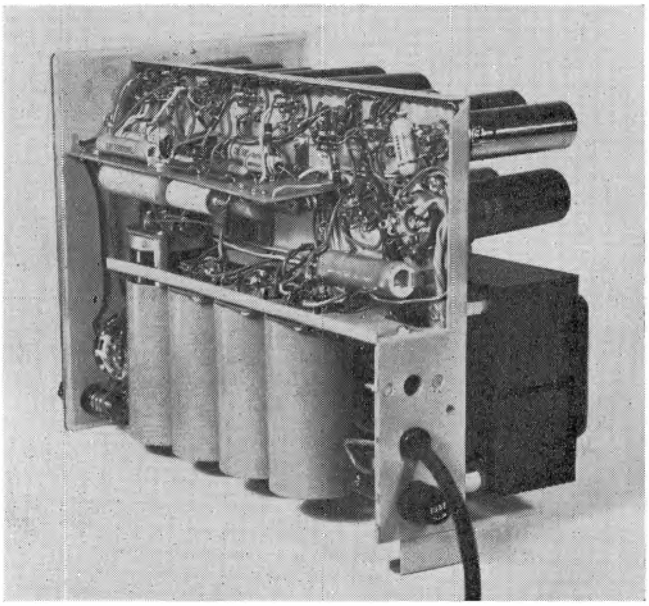


Fig. 10. An underside view of the generator

Construction

Final packaging takes advantage of the minimum number of components used. Case size is only 9½in high, 7in wide and 11½in deep. Total weight is 18lb. Controls are grouped on the panel for greatest readability and ease of manipulation compatible with size (see Figs. 9 and 10).

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Digital Computer Adding and Complementing Circuits

By C. D. Florida*

The design of transistor operated adding and complementing circuits is discussed with particular reference to d.c. coupled circuits suitable for use with double-gate shifting registers. Examples of circuits are given and illustrated by photographs of waveforms at a digit spacing of 5μsec.

MUCH of the equipment in a digital computer is repetitive, consisting of shifting registers, counters, gates and trigger circuits. While adding and complementing circuits can be built using assemblages of gates and trigger circuits, it can be more economical in components to design such circuits as complete entities. This has been the course followed in the design of the computer being built at the Electronics Laboratory of the Defence Research Board of Canada, and the results are presented in this article.

The computer under construction works in the serial mode and the design of the shifting registers is such that a digit is represented by a voltage level of either 0 volts (representing '0') or +10V (representing '1') which lasts for the duration of the digit period of 5μsec. If successive digits are the same the voltage remains constant so there are no waveform edges present either to define digit spacing or to initiate triggering. A consequence of this form of design is that either d.c. coupling has to be used throughout any subsequent circuits or sampling must take place each digit period.

The adding circuits described use d.c. coupling throughout while the complementer uses a mixture of d.c. coupling and sampling techniques.

Several different adding and complementing circuits have been devised and tested in the Laboratory. Of these, only two adding circuits and one complementing circuit will be described since it is considered that these circuits are improvements over earlier designs. Also, adding circuits can be built in two ways, using two half-adders or one full adder. Only the full adder forms will be dealt with as the derivation of the half adder form will be obvious.

Brief mention will be made at the beginning of the article of the arithmetic system of the computer and of the logical form of the adding and complementing circuits. This will be followed by a fairly detailed discussion of the two adding circuits and the complementing circuit.

The Computer Arithmetic System

Numbers are held in the computer in the form $a2^b$. That is to say, a semi-logarithmic binary system[1] is used. The semi-logarithmic feature need not concern us in this article but the fact that binary representation is used affects the design of adding and complementing circuits. Negative numbers, for instance, $-a2^b$, are held in two's complement form[2] and a few words describing this representation will be necessary in order that the functioning of the complementer can be understood.

The Two's Complement Representation of Negative Numbers

The two's complement of a number can be defined as that number which when added to the original number, gives an answer equal to zero. In the decimal system the same definition will hold if the word 'two' is altered to 'ten'. Thus the tens complement of 057 would be 943 since

$$\begin{array}{r} |\,057+ \\ |\,943 \\ \hline 1\,|\,000 \end{array}$$

The carry digit would be discarded as any machine has only a finite digit capacity. The combination of digits 943 can be regarded as a code signifying -057, and can be recognized by the occurrence of a 9 in the most significant digit position.

Similarly, in the binary system the two's complement of 01101 is 10011 since

$$\begin{array}{r} |\,01101+ \\ |\,10011 \\ \hline 1\,|\,00000 \end{array}$$

Again, 10011 can be regarded as a code for -01101 and can be recognized by the occurrence of a '1' in the most significant, or 'sign', digit position.

The two's complement of a binary number can be obtained in three different ways. First it can be obtained by subtracting that number from zero, and discarding the final borrow. For example:

$$\begin{array}{r} 00000- \\ 01101 \\ \hline 10011 \end{array} \quad \text{or} \quad \begin{array}{r} 00000- \\ 10011 \\ \hline 01101 \end{array}$$

Secondly, the number can be scanned, least significant digit first and the digits altered according to the following rules:

Initial '0's ...... leave unchanged
First '1's .... leave unchanged
Subsequent '0's .. change to '1's
Subsequent '1's .. change to '0's

Thus the number 0111001000, if scanned from the right-hand end would yield the same digits in the four least significant positions, made up of three initial '0's and the first '1'. All the remaining digits would be changed, '0's to '1's and '1's to '0's to give 1000111000.

The final way of conversion is to change all the digits from '1's to '0's and '0's to '1's and then to add a '1' in the least significant position. Examples are:

Number	01101	0111001000
Change	10010	1000110111
Add	1	1
2's complement	10011	1000111000

The reasons for the adoption of the two's complement representation of negative numbers have been fully discussed elsewhere[2] so will not be mentioned further here. It should be stated, however, that one important property of the system is that when performing addition no prior knowledge of the sign of the number is required, so that in a serial machine addition can proceed digit by digit as

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two numbers—no matter whether they are true numbers or complements—are fed into an adding unit.

Logical Forms of Adding and Complementing Circuits

Tables, sometimes called 'truth tables' may be constructed showing what outputs must be obtained from adding or complementing circuits when various digits exist at the inputs. From these tables it is usually possible to construct a schematic diagram showing a combination of gates, invertors, etc., which will obey the rules of the tables.

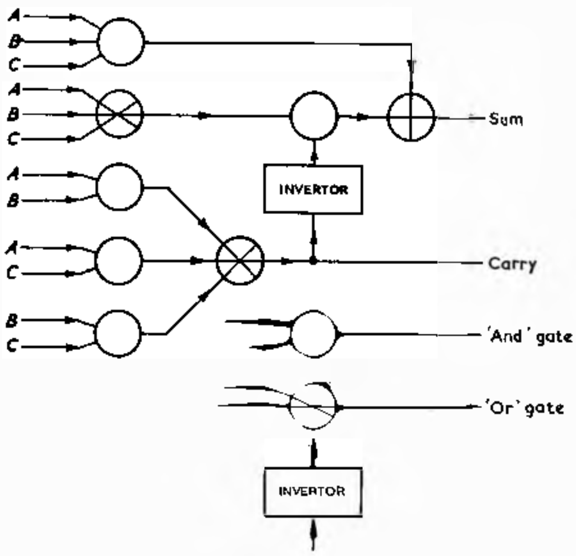

Fig. 1. Adding circuit logical diagram

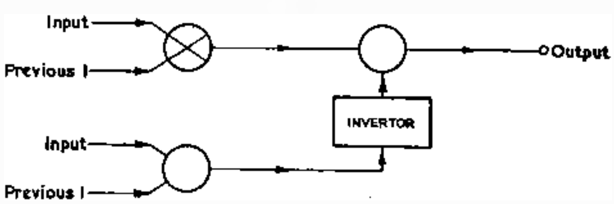

Fig. 2. Similar '0' dissimilar '1', circuit

Such diagrams are often termed logical diagrams since the gates, invertors, etc., perform the logical functions of 'and', 'or', 'not', etc.

Adding Circuits

A full adder circuit has three inputs, consisting of the two digits *A* and *B* being added during the current digit period and a further digit *C* which is a '1' if there was a carry from the previous digit period and which is otherwise a '0'. The table for such a circuit is shown below

A	*B*	*C*	SUM	CARRY
0	0	0	0	0
1	0	0		
0	1	0	1	0
0	0	1		
1	1	0		
0	1	1	0	1
1	0	1		
1	1	1	1	1

The requirements can be expressed in words as follows. There shall be a sum output when either *A or B or C* is a '1' or when *A and B and C* are all '1's but *not* when either *A and B* but *not C* or *A and C* but *not B* or *B and C* but *not A* are '1's. There shall be a carry output when either *A and B or A and C or B and C* and '1's.

A diagram showing components and their interconnexions which will fulfil these requirements is given in Fig. 1.

Dealing first with the carry circuit, it can be seen that if two or more of the inputs are registering '1's, at least one of the 'and' gates will register a '1'. Consequently the output of the 'or' gate and hence the carry output terminal will also be in the '1' state.

The sum circuit is a little more complex. If either *A* or *B* or *C* is a '1' the input triple-or gate will have a '1' as its output. If only one of the inputs is a '1' there will be no carry so the invertor output will be a '1'. This enables the '*and*' gate attached to it to pass the '1' from the input triple-or gate to the sum output line. If however there is a carry this path is blocked since the invertor output will be a '0' so that if a pair of inputs register '1's the sum output will be a '0'. Finally if all three inputs are '1's another path is opened via the input triple-and gate, the sum output 'or' gate to the sum output line, so that in this event the output will be a '1'.

Complementing Circuits

It was stated previously that there are three ways of forming the complement of a number. In this section the second way—that of scanning the number from the least significant end—will be used.

A complementer has one input and one output. In the scanning method of forming the complement there has to be another effective input which takes the value '0' until the first '1' has passed through unchanged, after which it assumes the value '1'. In the table which follows this pseudo-input will be termed 'Previous 1'. Also when the complementer is in use it is very convenient to be able to leave it in circuit at all times, arranging for it either to pass the number unchanged or to form the complement under the control of a second pseudo-input which will be labelled 'set'.

The table showing the requirements of a complementing circuit then appears as shown below.

INPUT	SET	PREVIOUS 1	OUTPUT
0	0	0	0
0	0	1	
1	0	0	1
1	0	1	
0	1	0	0
1	1	1	
1	1	0	1
0	1	1	

This table can be expressed in words very simply. If 'set' is '0' the input must be repeated at the output. If 'set' is '1' and the input and 'previous 1' are similar the output shall be a '0'. If 'set' is '1' and the input and 'previous 1' are dissimilar the output shall be a '1'.

A combination of gates and an invertor which obeys the rule "similar in '0'" out; dissimilar in "'1' out" is shown in Fig. 2 and will be recognized by those conversant with logical diagrams as a half-adder.

The information that the first '1' has passed through the complementing circuit needs to be retained until the end of the word. A trigger circuit, reset to '0' by a waveform occurring coincidently with the start of the word, is suitable for this purpose. This trigger circuit must be set to the '1' state only when the set state is '1' and the complementer output (or input) has been in the '1' state for one digit period. Since the '1' state is a voltage level without necessarily a return to zero at the end of a digit period, sampling is necessary to define the time at which the trigger circuit shall set, for which purpose a clock pulse whose triggering edge is coincident with the end of each digit period is suitable. The 'previous 1' information can thus be obtained and held by a triple-and gate and a trigger circuit as shown in Fig. 3.

Figs. 2 and 3 can now be combined to form a complete complementer logical diagram as in Fig. 4.

The 'Invertor Plus And' Circuit

In the logical diagram of both adding and complementing circuits there appears an invertor whose output is fed to an 'and' gate. The output of the invertor must register '1' when its input is '0' and '0' when its input is '1'.

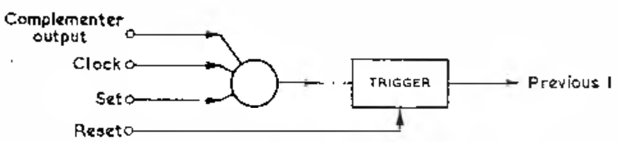

Fig. 3. Previous '1' circuit

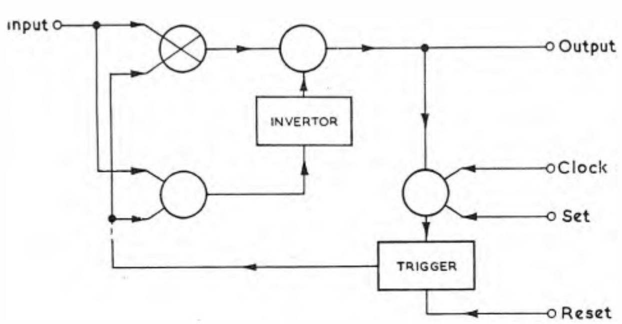

Fig. 4. Complementing circuit logical diagram

Throughout the computer '1' corresponds to $+10$V and '0' to zero volts. The problem is thus to design an invertor whose output will be $+10$V when its input is 0 volts and 0 volts when its input is $+10$V.

In the past, when the solution to this problem has been attempted using thermionic devices, the usual method has been to use an amplifier as an invertor and to attach a resistive potential divider to the output in order to attain the correct voltage levels. Such a technique can be continued when using transistors but leads to large voltage excursions across the transistor, high potentials and a need either for resistors of high stability or extra diode clamps.

Fortunately semiconductor research has provided the engineer with two additional devices with which to tackle this problem. One is the Zener diode and the other is the complementary transistor. Examples of invertor circuits using these aids are given in Figs. 5 and 6.

In the Zener diode circuits of Fig. 5 the invertor transistor is either off or is passing 5 volts/R_e amperes of emitter current according to the state of the input terminal. The resistor R_c is so chosen that in either case one of the collector catch diodes will be conducting. Current through the Zener diode in the reverse direction causes a potential difference of 15V to appear across it so that the correct output potentials of either 0 volts or $+10$V appear at the output terminal.

In Fig. 6 the transistor with the input joined to its base is the invertor. As in the previous circuits this transistor is either off or is passing a defined emitter current according to the state of the input. The resistor R_o is so chosen that the collector diode is conducting when the invertor transistor is passing current. Due to the relative potentials of this collector diode and the base of the complementary transistor the latter is cut off and has a reverse emitter to base potential of 2·5V when the invertor is on. Current through R_o, since it cannot in this event flow through the complementary transistor, passes through one of the output diodes, thereby defining one of the output potentials. When

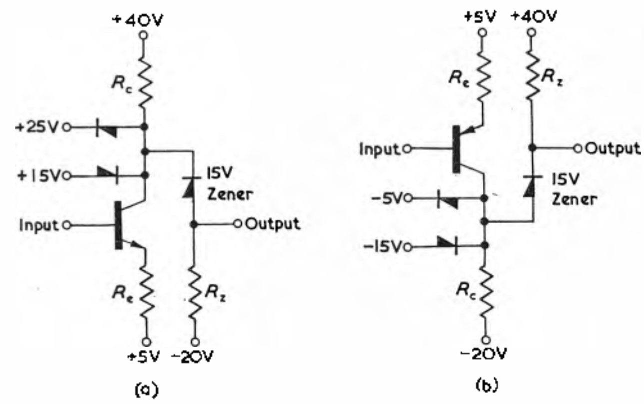

Fig. 5. Invertor circuits using Zener diode
(a) NPN transistor
(b) PNP transistor

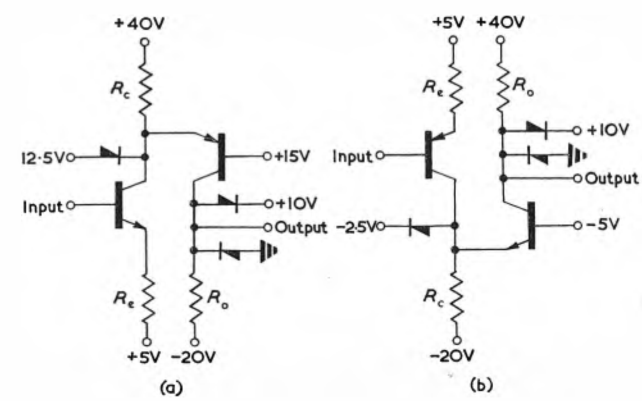

Fig. 6. Invertor circuits using complementary transistors

the input is in the other state the invertor is cut off and the current flowing through R_c becomes available to the complementary transistor. The resistor R_o is so chosen that under this condition the other output diode is conducting, thereby defining the second output potential.

Any of the circuits shown in Figs. 5 and 6 may have an 'and' gate attached to its output terminal so that it will behave as an 'invertor plus and' circuit. However, in the circuits using complementary circuits (Fig. 6) the same end may be reached more economically by using one of the output catch diodes as part of the 'and' gate.

To see how this is possible, consider Fig. 7, which is Fig. 6(b) redrawn with one change. This change is that the output diode which went formerly to a potential of $+10$V is now taken to a terminal marked input 2. It is assumed that this second input can assume the potential of either 0 volts or $+10$V and that it can accept the current supplied by R_o when in either state.

If input 1 is at $+10$V the complementary transistor is conducting and the output is clamped by diode action at 0 volts. It is then immaterial which state input 2 attains. If, however, input 1 is at 0 volts the complementary transistor is cut off and the current through R_o flows via the diode into the input 2 terminal. The output electrode thus takes

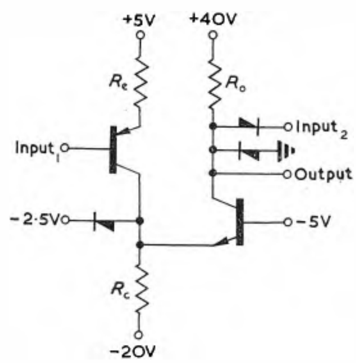

Fig. 7. 'Invertor plus and' circuit

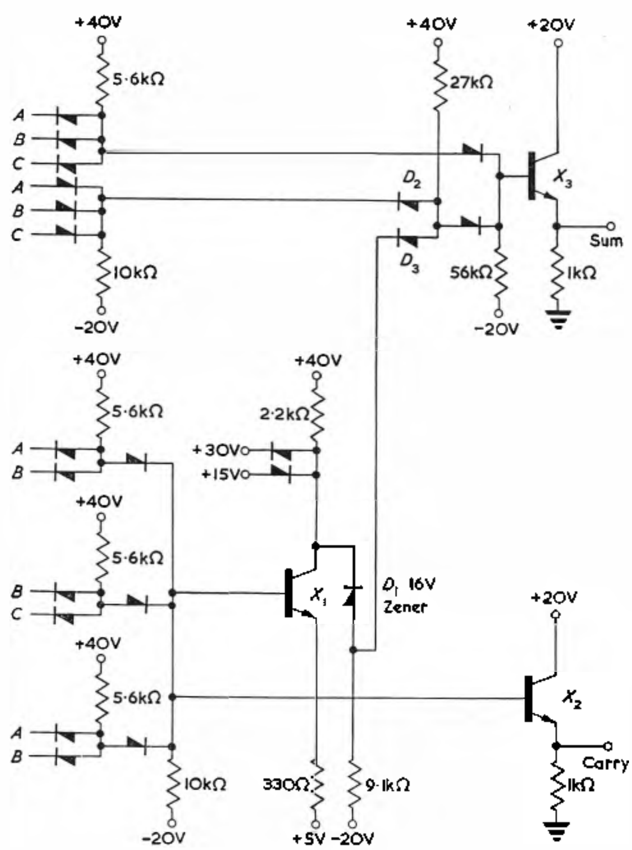

Fig. 8. Adding circuit using Zener diode

the potential of input 2. The circuit is thus obeying the following table, which is that of an invertor followed by an 'and' gate.

Input 1	*Input 1 inverted*	*Input 2*	*Output*
0 (0 volts)	1	0	0
0	1	1	1
1 (+10V)	0	0	0
1	0	1	0

In certain cases the other output diode, that returned to 0 volts in Fig. 7, can be used as part of an 'or' gate, thus effecting further economy. It is used in this manner in the second of the adding circuits to be described later, and further discussion will be withheld until then.

Practical Forms of Adding and Complementing Circuits

Two adding and one complementing circuit will now be discussed. The first adding circuit makes use of the Zener diode technique to attain the correct voltage levels, while the second adding circuit and complementer use complementary transistors to perform the same function.

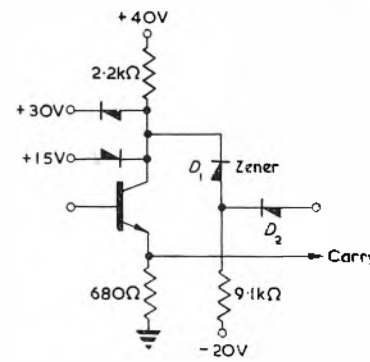

Fig. 9. Use of one transistor for inversion and impedance transformation

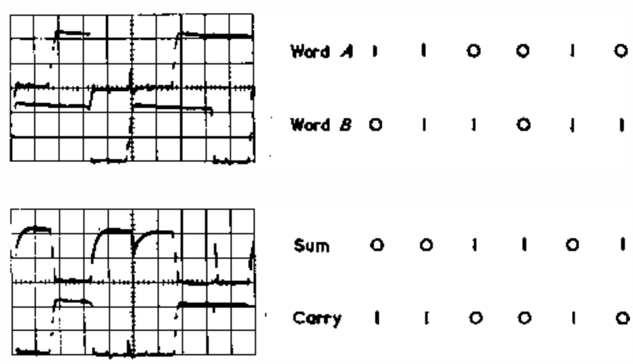

Fig. 10. Waveforms from an adding circuit using Zener diode

It will probably be obvious from what has been said before that in any of the circuits npn transistors can if desired be substituted for pnp and vice versa if the appropriate alterations in voltages, diode connexions, etc. are made.

The logical diagrams given previously assume that the output impedances are compatible with the input. This is not usually the case after a practical implementation of the diagram has been made, and in the circuits which follow the remedy adopted is the use of emitter-followers in the output stages.

An Adding Circuit Using A Zener Diode

The complete circuit diagram is given in Fig. 8. If this is compared with the logical diagram of Fig. 1 it will be seen that the gates on the left of each diagram are direct translations one of the other. The carry circuit, comprising the three two-input 'and' gates, the succeeding triple input 'or' gate and the emitter-follower X_2 can also readily be compared. The transistor X_1 operates as an invertor and the potential of its collector is constrained by the pair of diodes to lie at either +30V or +15V. The effect of the Zener diode D_1 is to reduce this potential to either +14V or −1V at the junction D_1D_3.

The potential at this junction will be at -1V only when there is a carry. Applied to the 'and' gate formed by D_2D_3 it serves to suppress the '1' which would otherwise be present. The output of this 'and' gate is thus '1' when either A, B or C is a '1' but is '0' when all inputs are at either 0 or 1 and when any two inputs are '1's.

The triple-input 'and' gate at the top left of the diagram rises to $+10$V only when all three inputs are at $+10$V. The output of this gate is passed via the 'or' gate D_4D_5 to the emitter-follower X_3 in a manner equivalent to that in the logical diagram.

The functions of both the invertor and the carry emitter-follower can be performed using one transistor only. A modification to Fig. 8 showing this is given in Fig. 9. A

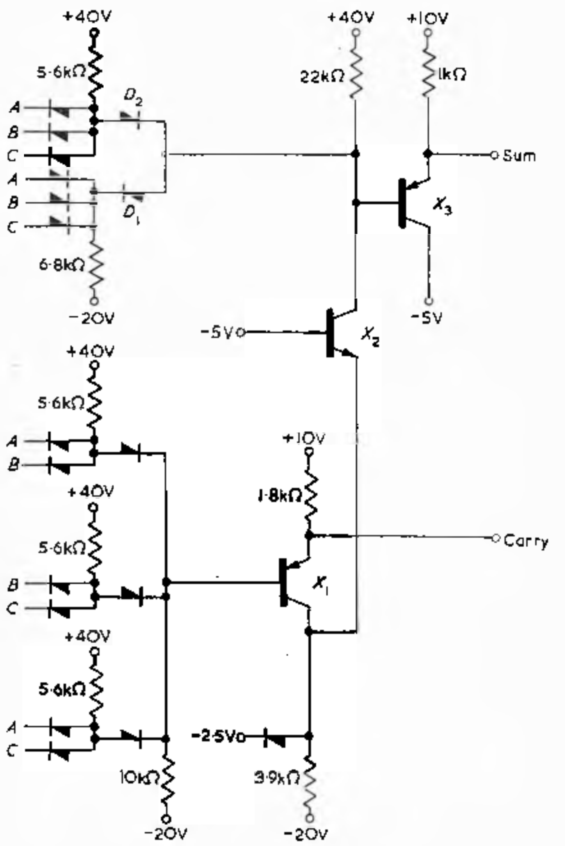

Fig. 11. Adding circuit using a complementary transistor

circuit with this modification is remarkable for its economy in transistors. Waveforms obtained from an experimental model using 3·5Mc/s transistors and a digit duration of 5μsec are shown in Fig. 10*.

An Adding Circuit Using A Complementary Transistor

The full circuit diagram is given in Fig. 11. The gates on the left of the diagram, to which the inputs are attached, are identical in form with those of Fig. 8, as also is the 'or' gate whose output is connected to the base of X_1. The output of this 'or' gate forms the carry, since it is at the potential corresponding to '1' whenever any two or all three inputs are '1's. The carry output is obtained at low impedance from the emitter of X_1.

Transistors X_1 and X_2 are arranged as an 'invertor plus and' circuit as shown in Fig. 7, the second input being provided by the output of the triple input 'or' gate to which D_1 of Fig. 11 is attached. Thus if there is no carry, X_1 will be conducting and X_2 cut off, so the potential of the collector of X_2 will be determined by the potential of the cathode of D_1: if all inputs are at '0' then the collector will be at '0', while if one input rises to the '1' state, the collector will also rise.

It was mentioned previously that in certain cases the diode of Fig. 7 shown returned to 0 volts can be used as part of a gate. Its use in this way is illustrated in Fig. 11 where the diode, D_2, is joined to the output of the triple input 'and' gate. The diode operates when there is a carry, so that X_1 is cut off, and X_2 conducting. If the carry is caused by two inputs being at $+10$V (in the '1' state) the output of the triple input 'and' gate will be at 0 volts. X_2 is arranged to draw more current than can be supplied by its collector load resistor so it obtains the balance via D_2, thereby ensuring that the collector potential is also near 0 volts. When, however, the carry is caused by all

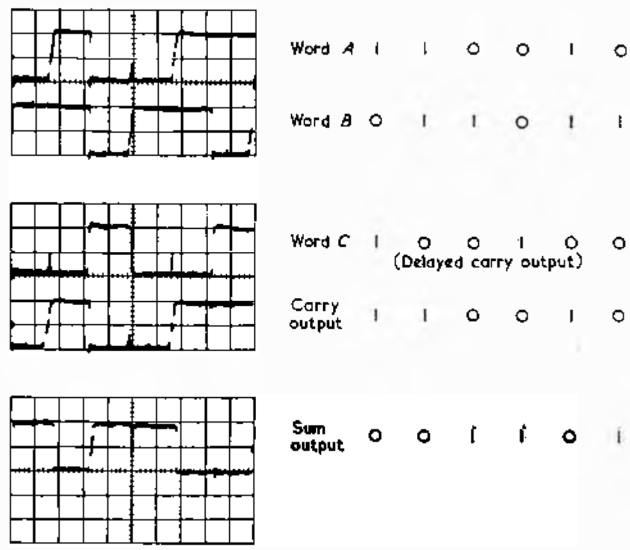

Fig. 12. Waveforms from the circuit of Fig. 11

three inputs being at $+10$V the output of the triple input 'and' gate, and consequently the collector potential of X_2, will also rise to $+10$V.

Waveforms obtained from a circuit as described using 3·5Mc/s transistors and a digit duration of 5μsec are shown in Fig. 12.

A Complementing Circuit Using A Complementary Transistor

The circuit diagram shown in Fig. 13 has been laid out in a similar manner to Fig. 4 so that the resemblance between the logical diagram and its practical embodiment may be revealed. The trigger circuit, enclosed within dotted lines, will not be described here as it is fully described elsewhere[3]. It is sufficient to say that a negative-going waveform applied to the capacitor attached to X_6 base will switch the trigger to the off, or '1' state, while a similar waveform applied to the other capacitor will switch the trigger to the on or '0' state.

The development of the remainder of the circuit follows naturally from what has already been discussed. Transistors X_1 and X_2, together with their associated components, form an 'invertor plus and' circuit identical with that shown in Fig. 7, while X_3 is an emitter-follower.

An Adder/Subtractor/Complementer Circuit

As an example of the more complex use of the techniques outlined in this article, Fig. 14 shows a combined adding and subtracting circuit. Use is made of the fact that when

* *In this and succeeding waveform diagrams the most significant digit of the word appears on the right-hand side of the waveforms: the written words appear in the normal manner with the most significant digit on the left.*

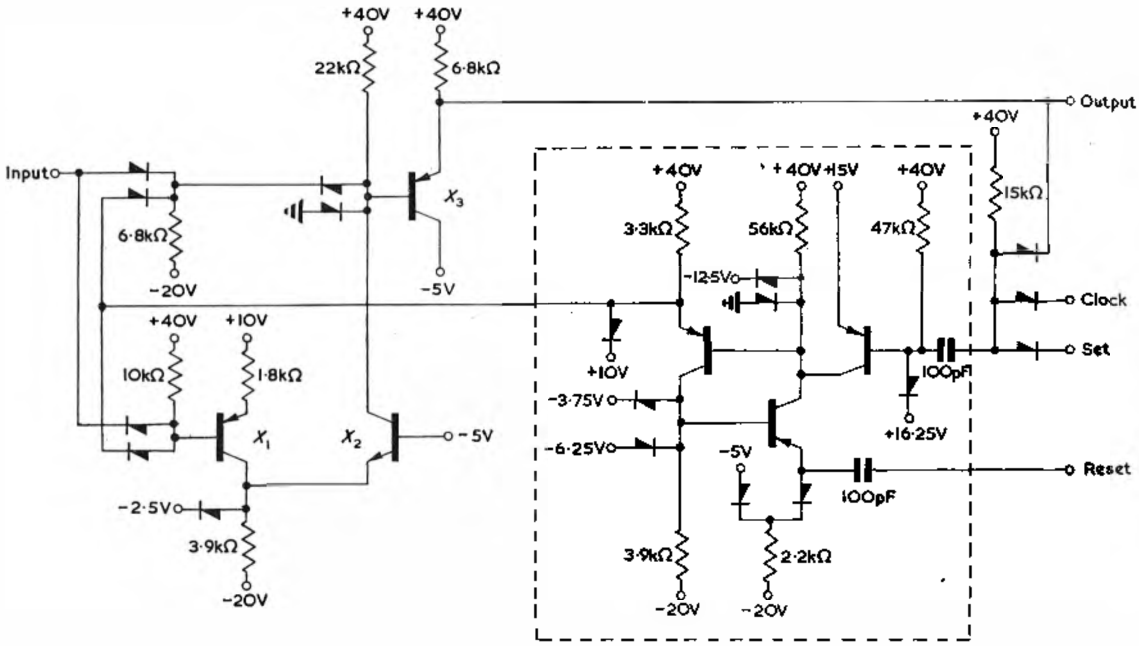

Fig. 13 (above). Complementing circuit

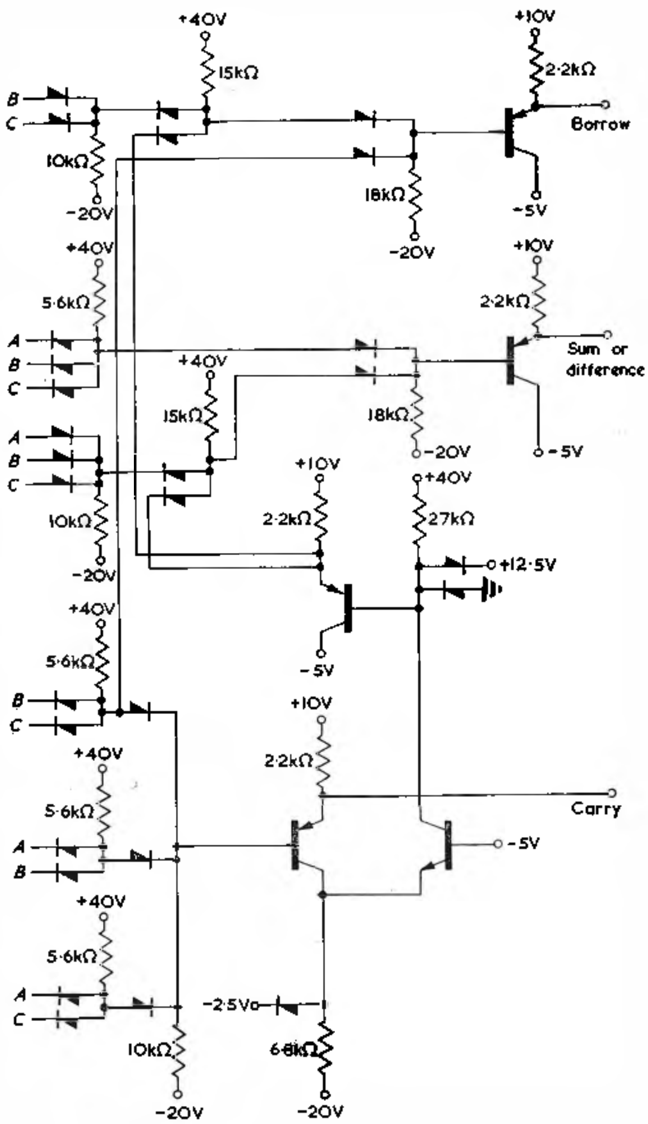

Fig. 14 (left). Adding and subtracting circuit

using two's complement representation of negative binary numbers, the sum or difference digits are both given by the expression*

$$S = (A + B + C)\,(\overline{BC + AC + AB}) + ABC$$

Here A is the addend or minuend, B the augend or subtrahend, C the previous carry or previous borrow and S the sum or difference digit. The borrow digit is given by

$$\text{BORROW} = (B + C)\,\overline{(BC + AC + AB)} + BC$$

and the carry digit by

$$\text{CARRY} = BC + AC + AB$$

The circuit will not be described in detail since it follows immediately from preceding descriptions and the equations.

When the circuit is used for addition terminal C is held by a trigger circuit at the potential obtained at the carry output during the preceding digit period. To change to subtraction the trigger inputs are switched from the carry output to the borrow output.

It was shown earlier that the two's complement of a number may be obtained by subtracting that number from zero. Consequently if the circuit is arranged as a subtractor and input A is held at zero the two's complement of a number entered at B is obtained at the difference output.

Waveforms obtained from this circuit are shown in Fig. 15.

Conclusions

The manner in which adding and complementing circuits have been designed for use in the D.R.B. computer has been discussed and full circuits presented. These circuits

* *This is read as: Sum Digit is '1' when either (A OR B OR C BUT NOT B AND C OR A AND C OR A AND B) OR A AND B AND C ARE '1's. The OR function is signified by an arithmetic plus sign, the AND by an algebraic multiplying command, and negation by the horizontal bar.*

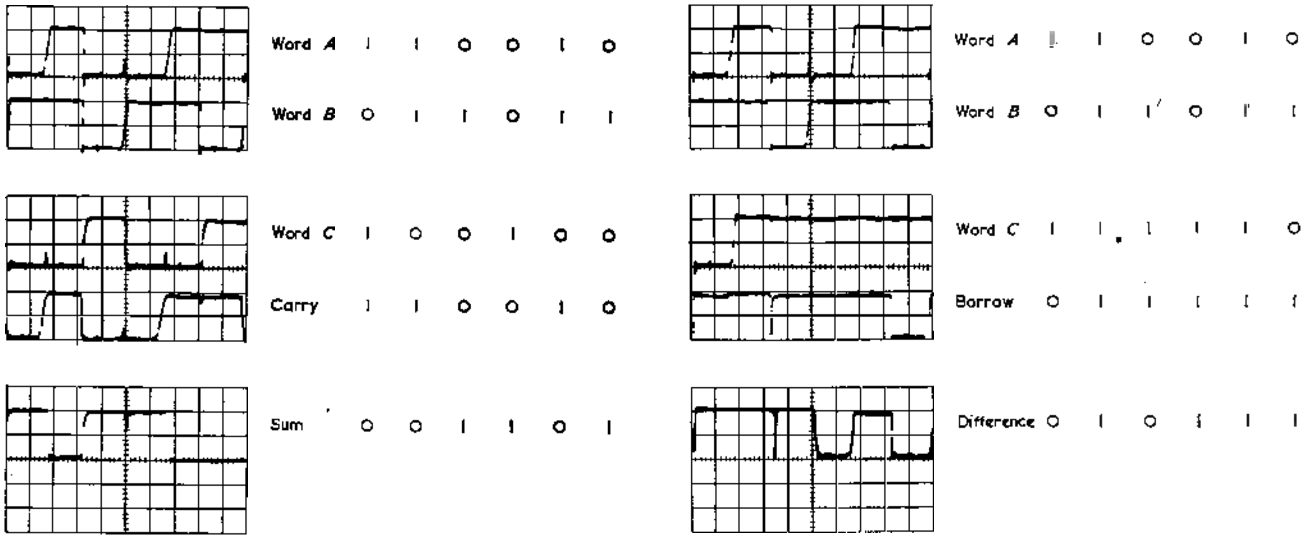

Fig. 15. Adder/subtractor waveforms

prove to be simple and reliable. Their performance in the way of speed is shown by the waveform diagrams and can be seen to be faster than necessary. An improvement in the economy of the circuits is therefore possible by the reduction of currents.

Acknowledgments

The experimental work on the circuits described in this article was performed by Mr. G. Mackie and Mr. R. R. Cretchley. Earlier development work was carried out by Mr. A. Poznanski of Canadian Marconi Company and by Mr. J. Rywak of Northern Electric Company Ltd.

The author wishes to thank the Defence Research Board of Canada for permission to publish this article.

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A Conditional Probability Computer

Among the exhibits shown for the first time at the recent Open Day at the National Physical Laboratory at Teddington was a conditional probability computer.

This machine which is in the experimental stage, is described as a machine which can learn from past experience.

The computer has five input channels labelled j, k, l, m, n. It consists of 31 similar units which accumulate statistical information concerning the patterns of input activity presented in these channels. When some statistical data has been accumulated, and when one or a group of the input channels is activated, the computer determines, on the basis of its stored statistical information, whether this group is usually accompanied by activity in another channel or channels. If it is, the computer makes an inference of activity in the other channels (shown by red illumination of the panel above the computer). The precise conditions for an inference are described below.

The 31 units are essentially counters; five of them count the numbers of occurrences of activity in the respective input chanels, while a further ten units count the numbers of occurrences of simultaneous activity in each of the ten possible pairs of input channels, and so on for groups 3, 4 and 5 channels. The counts are represented in the units by the amount of charge on a capacitor, which is deliberately made to leak towards the level corresponding to zero count. The introduction of leakage ensures that the inferences made by the computer are governed more by recent events than by less recent events.

The conditional probability of activity in the l channel, given that there is activity in the j and k channels, is given by:

$$\text{Probability of } l \text{, given } j \text{ and } k = p_{jk}(l)$$

$$= \frac{\text{count stored in } (jkl) \text{ unit}}{\text{count stored in } (jk) \text{ unit}}$$

When the j and k inputs of the computer are activated simultaneously, it computes the quantities $p_{jk}(l)$, $p_{jk}(m)$ and $p_{jk}(n)$. If any of these exceeds a predetermined threshold value, an inference is made of activity in the l, m or n channel.

The computer constitutes a 'learning machine' with many possible applications, and was demonstrated controlling a device which follows the boundary between a black and a white area, irrespective of whether black is to the left and white to the right or vice versa. The application of these principles to industrial control will involve greater complexity but it appears to be possible to make applications and work on them is in progress.

Demonstration of a Civil Doppler Navigation System

A Viking aircraft of Airwork Services Ltd has recently been fitted with a Marconi Type AD 2300 Doppler Navigation system adapted for use on civil airlines and has just completed a demonstration flight in Europe.

In a further demonstration flight at Blackbushe the aircraft flew on a triangular course of some 120 miles during which the general performance of the navigation system was observed.

The particular points of interest were as follows:—

Automatic signal acquisition. This is of importance at the point shortly after take-off. No action is required by the aircrew except to switch on the equipment.

Accuracy. The navigation equipment included a computer with outputs of Northings and Eastings in nautical miles referred to the readings at the start of the triangular course. Readings of Northings and Eastings at the start and finish of the course were observed.

Performance in Turns. Full turns of 360° and bankings of 450° were carried out to demonstrate change of heading, drift and ground speed.

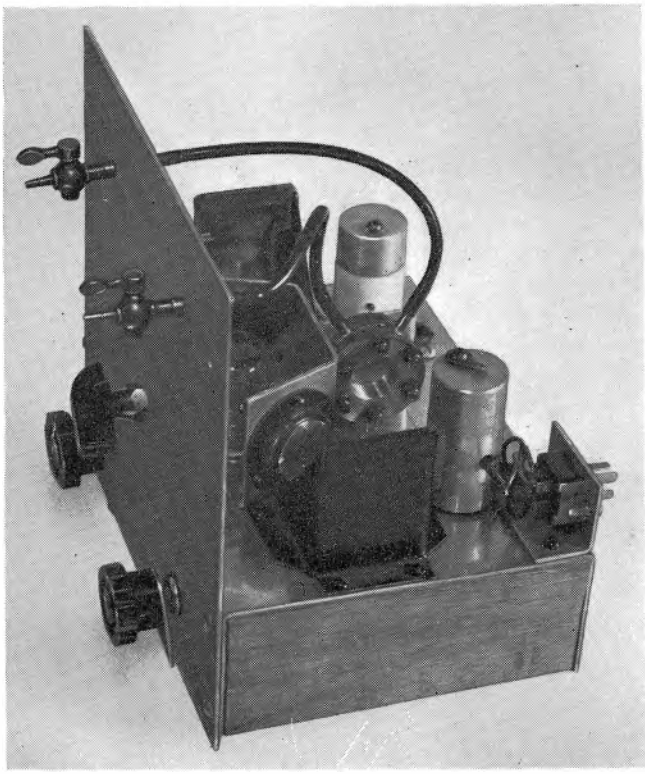
Performance in Climb and Descent. During the climb at take off and the descent during approach the artificial horizon showed the aircraft altitude, while the aerial pitch stabilization was seen in operation. Climbs and descents at a height of about 6 000ft were also made at the rate of 1 000ft/min.

A Sensitive Defocusing Photo-electric Pressure Transducer

By J. R. Greer*, B.Sc., Ph.D.

A design for what is believed to be a novel type of photo-electric pressure transducer using germanium photocells is described, and it is shown that, as a result of very high optical efficiency, symmetrical outputs of over 20V may be obtained for pressure differences of a few centimetres of water. The transducer is both mechanically and pneumatically robust, and is stable over long periods.

THE optical lever type of pressure transducer, despite its inherent simplicity of operation, usually suffers from the disadvantage of requiring frequent alignment, and of giving a low output. Moreover, it is unsatisfactory for use when both negative and positive pressures are to be recorded since, while the zero can normally be offset to some arbitrary value to enable negative pressures to be measured, it is an inevitable consequence of this that any alteration to gain will affect the position of the zero. The defocusing manometer was designed to take full advantage of the new miniature transistor-type photocells, and has proved a mechanically and pneumatically robust unit which gives an output (prior to amplification) of ± 20V for an applied pressure of ± 5cm of water. As the instrument, by its nature, operates symmetrically about zero, alteration of the gain does not affect the position of zero and positive or negative pressures may be traced with equal ease. It has thus proved of considerable value in respiratory research, both for pressure and for flow measurements. The manometer is normally used with gas as coupling medium as this has been found much more satisfactory for respiratory investigations than fluid transmission; but fluid pressures may be measured by interposing a small bottle, similar in principle to an aspirator, as a fluid trap. Alternatively both sides of the diaphragm assembly may be filled with transparent fluid, two additional taps being added for the purpose.



The complete instrument

Principle of Operation

If a thin elastic circular diaphragm is mounted under tension, and a small uniform pressure is applied to one side of it, the lamina will assume a shape approximating to that of a sphere whose radius of curvature, provided it remains large, will vary inversely with the applied pressure. If, therefore, such a lamina has a highly reflective surface, and a beam of parallel light impinges on it, the reflected rays may be parallel, divergent or convergent, depending on whether the pressure behind the lamina is zero, positive or negative respectively. Since the sensitive part of a germanium photocell is less than 1mm², the changes in image size resulting from changing pressure appear as corresponding changes in intensity, provided that the centre of the cell is near the optic axis.

Practical Design

The first problem to be solved was that of choice of material for the lamina or diaphragm. This must be thin, elastic, and with a bright surface. Aluminium and duralumin (each ·001in) are adequate optically, but aluminium proved insufficiently elastic, while duralumin began to show signs of corrosion from expired moisture.

A diaphragm of silver foil, about ·001in thick, has been in continuous service for over a year now, and has proved most satisfactory, having shown no tendency to fatigue or tarnish; it is however, difficult to obtain. Recent experiments have used thin Terylene foil (I.C.I. Melinex) on to which a thin layer of aluminium had been evaporated in vacuo. This gives a thirty-fold increase in sensitivity, with no apparent disadvantages.

The detail of the diaphragm assembly is shown in Fig. 1. Good machining is essential on the two inner faces, to ensure that the diaphragm is not cut by the tensioning lip and groove. In assembly all faces are smeared with petroleum jelly or vacuum grease, and the plastic or silver cut roughly to size and laid in position. The brass faces are then aligned and clamped lightly while holes are burnt or punched as appropriate in the diaphragm, and screws are inserted and tightened in rotation. No separate tensioning of the diaphragm is necessary.

The range of linear operation is considerably increased by using a symmetrical layout and reflecting light from both sides of the diaphragm on to two cells connected

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in push-pull (Figs. 2 and 3). Variations in lamp brightness or h.t. voltage affect both cells equally, and therefore tend to be balanced out; while thermal effects are minimized by mounting the cells in aluminium pillars which have large thermal inertia and are thermally coupled through the chassis.

Component layout in general is not critical, but the chassis should be sufficiently rigid to prevent vibration of the mirrors and other optical parts. Again, although total enclosure is unnecessary, zero drift is reduced if the cells are protected from draughts or strong lights.

Photocells and Lamp

For maximum voltage output the most suitable cell is the STC type P50A, (the lower working voltage of the Mullard OCP71 off-setting its greater current sensitivity). These cells vary appreciably in sensitivity and also in dark current; normally it is convenient to make the loads and applied voltages equal (Fig. 3(a) and (b)), apart from a small fine set zero control, and compensate for different sensitivity by reducing, with a simple shutter, the light applied to one cell or the other (Fig. 2(a)). If, however, one cell has a substantially greater dark current than the other this difference may give rise to unequal sensitivity to positive and negative pressures and should be remedied by switching off the light and then balancing the output by shunting the better cell so that its 'dark current' is raised to equal that of its neighbour. (A 5MΩ potentiometer is useful.) This should not require alteration until the cells are replaced.

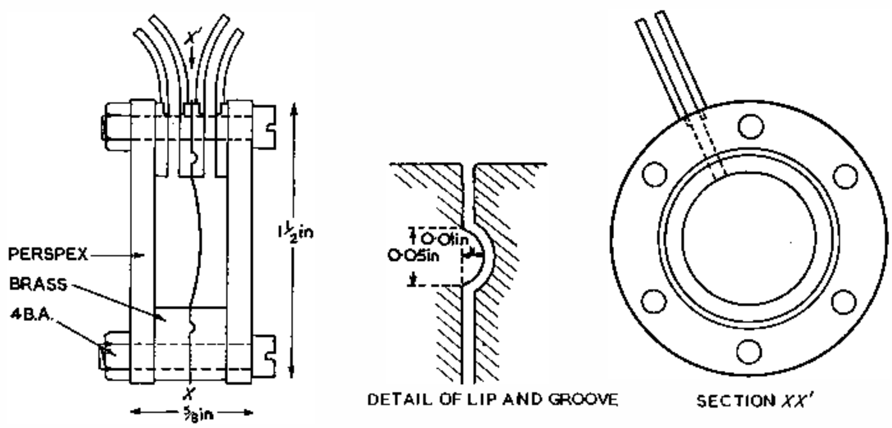


Fig. 1. Diaphragm assembly and detail of tensioning lip

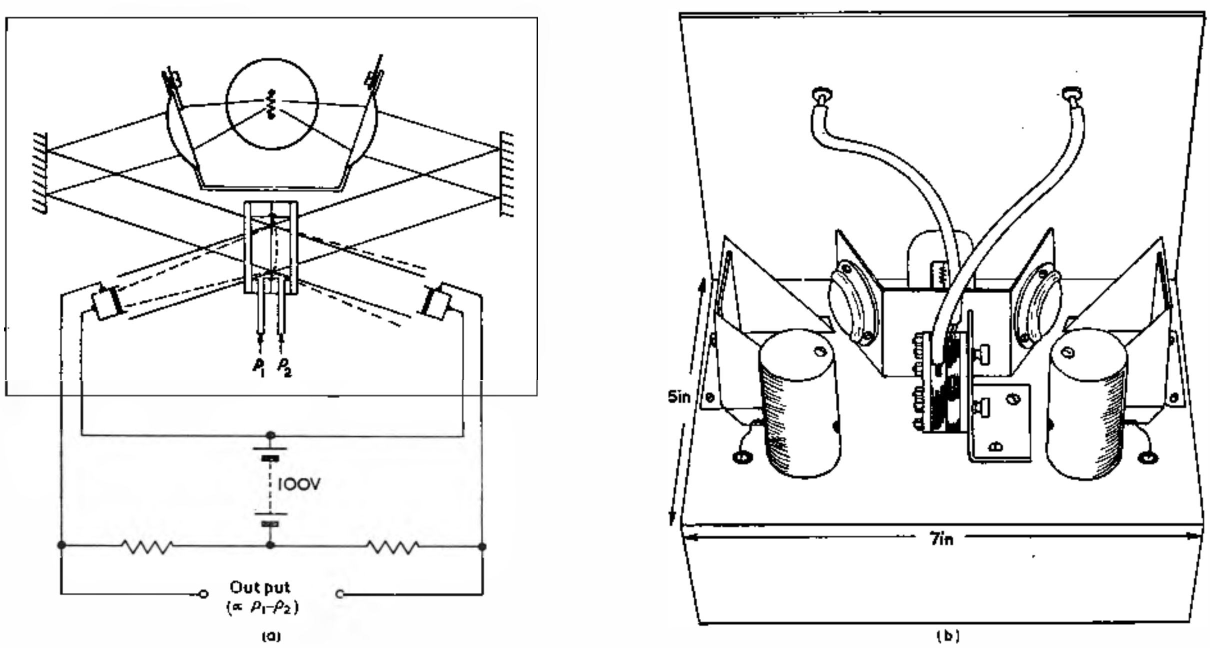


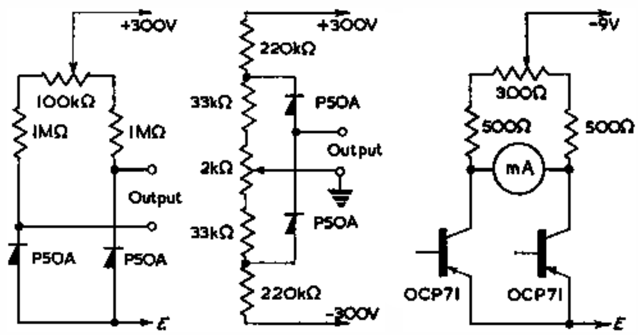
Fig. 2. General layout of components

In order to obtain a lamp with a thick, rigidly mounted filament which would have a long expectation of life a car headlamp bulb was chosen. Owing, however, to the sensitivity of the photocells and the high optical efficiency of the system, this provides much more light than is required, and it has been found convenient to run the bulb at half its rated voltage.

For battery operation, particularly with Mullard OCP71 cells, a very much smaller lamp would suffice, the miniature prefocus types with a straight filament supported at each end being especially suitable.

Pen Recorder Amplifiers

Fig. 4 shows the circuit of the associated pen recorder amplifier, from which also are derived the necessary power supplies. This amplifier (which has an effective slope of about 1mA/V working into a 3kΩ recorder) has been modified by the addition of a variable resistor that allows the cathode potential to be set to the mean potential of



(a) (b) (c)

Fig. 3. (a and b) Alternative circuits for balanced or single-ended output (c) Battery operated transducer

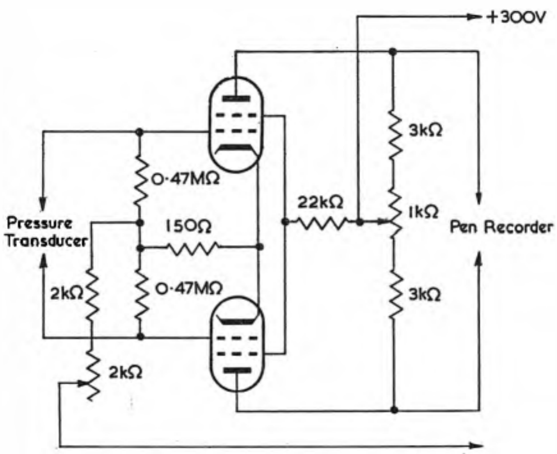


Fig. 4. Pen-recorder amplifier for use with circuit in Fig. 3(a)

the photocell anodes, using the circuit of Fig. 3(a). Fig. 3(b) may be more convenient if a single-ended output is required, as for a spirometer, and a negative line is available. The values shown are not critical, but are suitable for use with 300V positive and negative lines. Whatever supply is used, the potential across each cell, when they are not illuminated, should not exceed 100V.

Measurements of Small Pressure Gradients

Several authors (Pressey[1], Nisbet[2]) have discussed the method of determining the volume of air that a patient has inspired or expired over a given period by measuring the pressure gradient across a fine gauze screen. This pressure has been shown (e.g. Lilly[3]) to be proportional to rate of flow under certain conditions, so, after integration against time, a voltage may be derived whose amplitude is an index of inspired volume. The variations in flow rate observed during respiration may be very large, so the wide dynamic range of this manometer makes it eminently suitable for such measurements, and an electronic spirometer of this type, which is equally suitable for measurements on infants or adults has been designed and is now receiving clinical trials.

Battery Operation

The possibility of using a low-powered lamp, together with the Mullard OCP71 cells has already been referred to, and a typical circuit is shown in Fig. 3(c). Using the values shown pressures of ±50cm of water applied to a ·001in Terylene diaphragm gave an output of ±5·5mA: using a ·00025in diaphragm, this output would correspond to an applied pressure of 10 to 15cm of water. This current was measured on the 10mA range of an Avometer, and was reduced to about ±1mA when the meter was replaced by a pen recorder of 3kΩ internal resistance.

While detailed experiments have not yet been carried out, it will be clear from the foregoing that a pressure

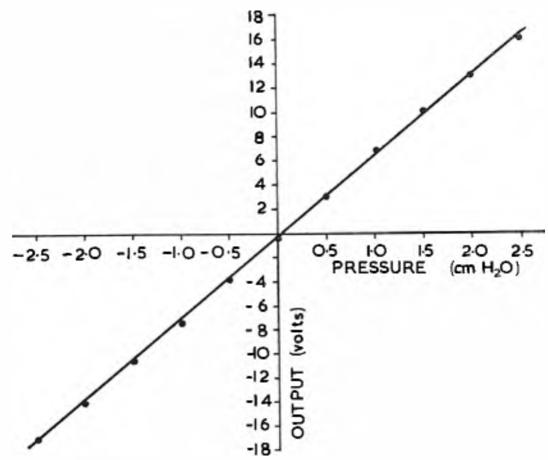


Fig. 5. Response of transducer with ·00025in Terylene diaphragm to positive and negative pressures

transducer requiring at most a 9V battery, and capable of operating a meter or pen recorder directly, could easily be constructed for any pressure range from 1cm of water upwards.

Sensitivity, Linearity, Drift and Dynamic Range

The sensitivity and linearity were measured with a calibrating water manometer (Greer[5]). The output was 1V per 10cm water pressure for a silver diaphragm, or 1V for 2mm water pressure with a ·00025in Terylene diaphragm. In each case the drift was less than ±100mV over several hours when d.c. connected, or less than 10mV if connected through a coupling with a 20sec time-constant. The linearity calibration is shown in Figs. 5 and 6. No destructive tests have yet been carried out to determine the breakdown pressure, but the unit with the silver diaphragm has been frequently cycled up to 250mm Hg (= 3 300cm H_2O) without any subsequent signs of fatigue, such as an erratic zero, so it would seem that a dynamic range of not less than 1 000:1 can reasonably be claimed. The unit with the 0·00025in Terylene diaphragm has also been tested up to 250mm Hg, and while this of course is well outside its linear range, its subsequent performance was unimpaired, even for pressures of less than a few millimetres of water.

Frequency Response and Hydraulic Admittance

The frequency response is governed largely by the

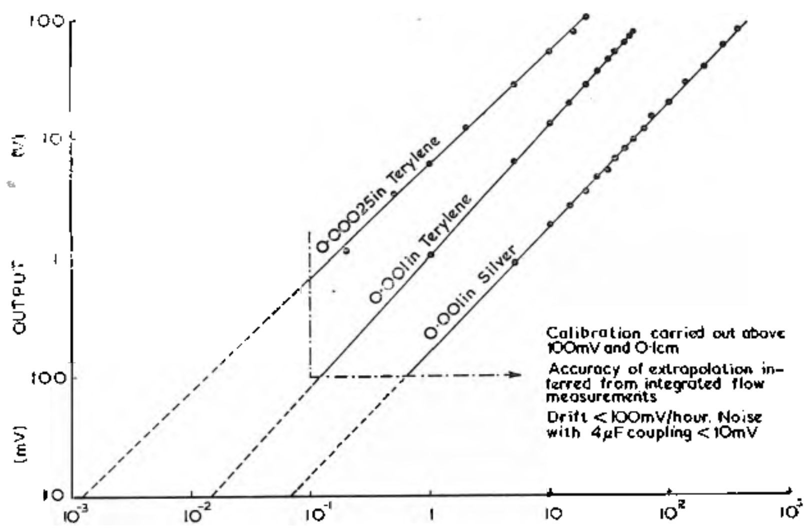


Fig. 6. Sensitivity of different diaphragms

cavity resonance of the diaphragm casing, and diameter of connecting tubing. In the present unit it extends to approximately 500c/s.

When a pressure gradient is applied to the diaphragm, its resultant radius of curvature (and hence its volume displacement) will depend on the thickness, tension, and elasticity of the diaphragm, as well as on the pressure applied: the output, on the other hand, depends only on the curvature of the diaphragm. For this reason it is convenient to quote the hydraulic admittance as being 1mm³ displacement per volt output; the corresponding pressure (for the diaphragms described) being obtained by reference to Fig. 6.

Acknowledgments

It is a pleasure to thank Professor Ian Donald of Glasgow University Department of Midwifery for his interest and advice on clinical requirements, Mr. J. Macfadyen of the Regional Physics Department who constructed the instrument, and Mr. J. T. Lloyd of the Glasgow University Department of Natural Philosophy who made available the aluminized Terylene foil. The work was carried out at the Glagow Regional Physics Department as part of a research programme on neonatal respiratory disorders under a grant from the Scottish Hospital Endowments Research Trust.

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A Mobile Medical Colour Television Unit

What is claimed to be the first mobile medical colour unit to be designed and manufactured in Great Britain was formally handed over recently by its makers the Marconi Wireless Telegraph Co. Ltd to Smith Kline and French Laboratories Ltd who commissioned its manufacture.

Last year Smith Kline and French sponsored a number of colour television demonstrations using a similar mobile unit borrowed from the United States. So much interest and enthusiasm was displayed in colour television as a medical teaching medium that the Company decided to provide a mobile unit to be based permanently in the United Kingdom.

Programmes this year have been arranged to be given at the British Medical Association's Annual Meeting at Birmingham and in London and Southampton, and at other medical meetings in the country.

The vehicle itself weighs 8 tons gross, 2 tons of which is accounted for by the colour television apparatus. It is divided into three sections, as under:—

Rear Compartment.—This houses the apparatus racks, including the encoding equipment which combines the red, green and blue signals from the camera into a composite signal of a type which can be carried, if required, over long distances via landline or microwave link. N.T.S.C.-type coding modified to British 405-line requirements is employed for use in this country, but the equipment is suitable for use with other line standards. The apparatus room compartment is normally staffed by one maintenance engineer.

Centre Compartment. This is the control room area and in it sit the technical director and vision mixer, two camera control operators and the sound mixer; the team working under the personal direction of the producer. In this room the respective outputs from the two colour cameras and the microphones are checked and integrated into the programme as seen and heard by the audience.

Three colour monitors, type BD.875, each employing a 21in three-gun shadow mask type of colour tube, are in use in the control room one for each camera channel and one acting as a line monitor. The black-and-white preview monitor uses an 8½in picture tube, while the two picture and waveform monitors each employ 14in picture tubes and 5in waveform tubes. One interesting feature is the arrangement whereby the colour monitors are underslung on a mono-rail system which enables them to be withdrawn into the control room proper for minor adjustments, to the forward compartment for major servicing or to be withdrawn completely if occasion demands.

Forward Compartment. This, in addition to containing the driver's seat and controls (right-hand drive) provides access to the back of the camera control equipment and is also designed to give a useful servicing area.

The colour cameras are Marconi 3-Tube Image Orthicon cameras, Type BD.848, the fundamental design of which is similar to that of the colour cameras supplied by Marconi's to the BBC for their experimental transmissions in colour, but incorporating various improvements in the light of subsequent research. Special features of the camera include finger tip control of the camera position, an independently inclined electronic view finder and an adjustable ventilator for additional cooling when the camera is working in usually high ambient temperatures. The cameras are designed to operate via a single 88-way camera cable, two 100ft lengths of which are carried on drums in the forward compartment of the vehicle, one end of each being permanently connected to the equipment. Two additional drums of 88-way cable, each carrying 250ft lengths are also carried for use when the camera has to be used even further away from the vehicle.

The main display unit is a Marconi colour television projection unit type BD.876. This provides a picture 8ft by 6ft which can be viewed comfortably by up to 300 people. A 21in colour television receiver manufactured by Murphy Radio Ltd is also provided as a second display unit, for use, if required, in an additional viewing room.

No power generating equipment is provided since the mobile unit will normally be operated at sites where power supplies are readily available.

The Measurement of Transistor Voltage-Current Characteristics Using Pulse Techniques

By B. J. Cooper*, B.Sc. (Eng.)

A method of measuring transistor voltage-current characteristics using pulse techniques is described. This gives ambient temperature characteristics and reduces the effects of junction heating and thermal runaway present in a d.c. measurement.

A circuit designed to investigate transistor characteristics using this technique is given, the voltage-current curves may be read from meters or displayed directly on an external oscilloscope. Typical results are given.

IN order to select junction transistors for some applications it is necessary to measure the maximum permissible collector-emitter voltage. It is difficult to measure this voltage by d.c. methods as the collector leakage current (I_{co}'), at high collector voltage causes internal heating of the transistor. This increases the collector current and in many cases the transistor may suffer permanent damage due to thermal runaway. Even if runaway does not occur, the increased collector current means that an ambient temperature characteristic cannot be obtained. If the transistor is fed with short duration voltage pulses, instead of d.c., then the power dissipated is small and the voltage and current pulse amplitudes will give the ambient temperature characteristic. This technique has been used for testing point-contact transistors[1] where the low thermal capacity of these devices necessitates the use of a low mark-to-space ratio. This limitation is not so severe with junction transistors and it is possible to use a higher mark-to-space ratio. Pulses of 2msec duration at a rate of 50 per second are convenient as 50c/s mains may be used to synchronize, and a pulse width of 2msec ensures that distortion of the current pulse through the transistor, due to hole storage effects and capacitance, is negligible. When fed with such pulses, having a mark-to-space ratio of 1:9, the dissipa-

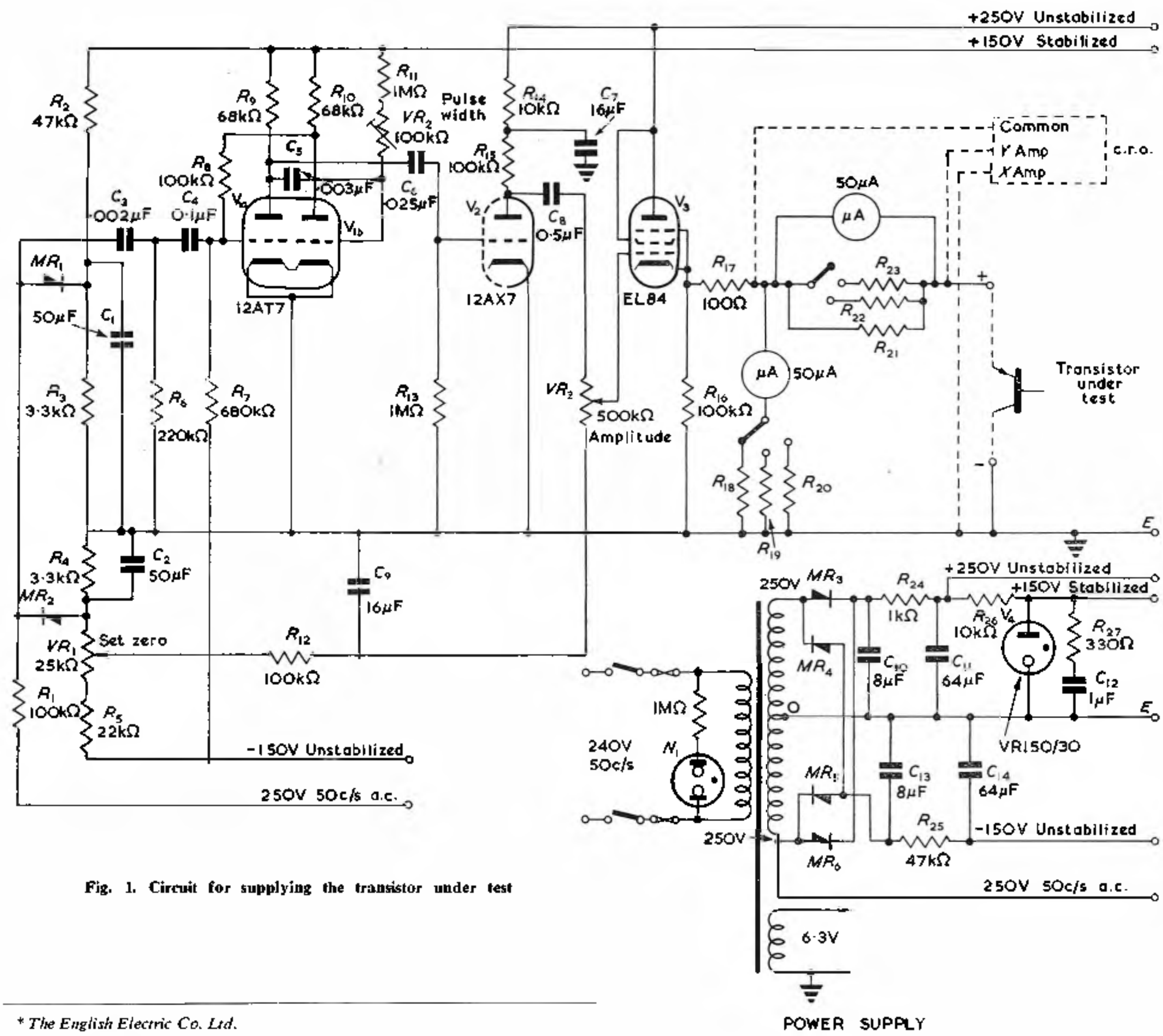

Fig. 1. Circuit for supplying the transistor under test

* The English Electric Co. Ltd.

tion of the transistor is one-tenth of that in a d.c. measurement. If the current and voltage be measured on moving-coil d.c. meters, which indicate average values, the meter readings will be one-tenth of the pulse amplitude in each case.

Circuit

A suitable circuit for supplying the transistor under test is shown in Fig. 1. Valves V_1 and V_2 form a monostable multivibrator which is triggered at mains frequency via clipping network R_1, MR_1, MR_2. This network is fed at 250V 50c/s and produces a square wave output of approximately 10V peak-to-peak amplitude. This is differentiated by C_3R_E and fed to the grid of V_1 to trigger

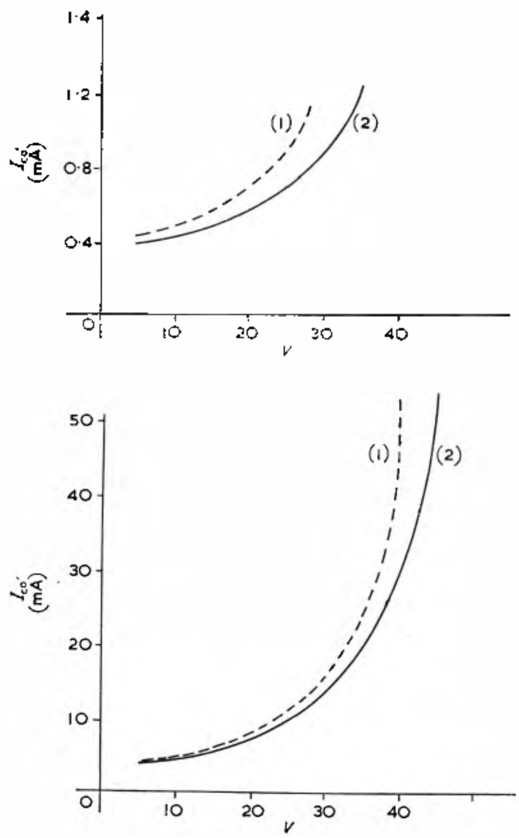

Fig. 2. Voltage-current characteristics of two power transistors. Curve (1) d.c. supply. Curve (2) pulse supply
(a) Transistor type V30/30P
(b) Transistor type 2N173

the multivibrator. The pulse width of the multivibrator is set by variable resistor VR_2 to be 2msec. The 150V h.t. supply to the multivibrator is neon stabilized to avoid pulse width variation. The output from the anode of V_1 is amplified by valve V_3 and fed to a cathode-follower output stage V_4 via the output level control VR_3. V_4 is held non-conducting by the negative potential set on potentiometer VR_1 and the application of positive going pulses to the grid cause it to conduct, giving positive going voltage pulses at the cathode, which are taken via the metering circuit to the transistor under test. Outputs proportional to voltage and current respectively may be fed to the X and Y amplifiers of a direct-coupled oscilloscope to give a direct display of the transistor characteristics. With the circuit as shown the maximum output voltage available is approximately 150V at zero current, the output impedance being approximately 500Ω with the meter switch set to the 100mA range. The meter calibration is such that equivalent d.c. values are read directly, and the following ranges are available:

Collector Voltage	*Collector Current*
0-1V	0-1mA
0-10V	0-10mA
0-100V	0-100mA

Applications

When used with a cathode-ray tube display the circuit enables transistors to be graded very rapidly. For more precise measurements, meter readings are plotted to give the voltage-current characteristics. The low power dissipation in the transistor under test permits the application of relatively high voltages without damage, and also enables reproducible measurements to be made of characteristics at voltages close to the Zener voltage. This would be impossible with conventional d.c. techniques. Collector-emitter voltage-current characteristics for two germanium power transistors are shown in Fig. 2. The full curves were obtained using the circuit described and the broken

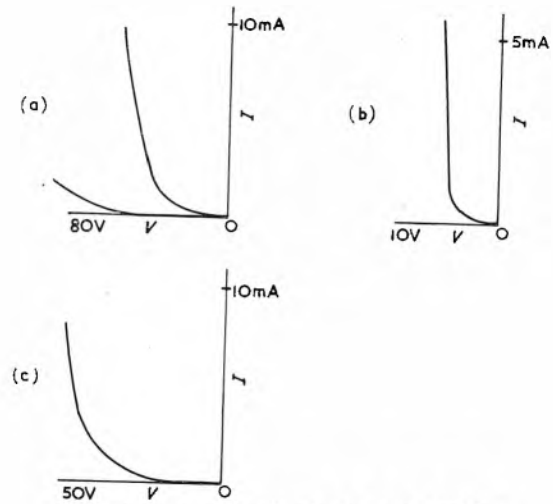

Fig. 3. Oscilloscope records using the pulse technique

curves show the results with a conventional d.c. measurement. In both cases it can be seen that the lower power dissipation in the pulse measurement gives a smaller junction temperature rise and hence a lower collector leakage current than in the d.c. case. Fig. 3 shows oscilloscopic records taken using the pulse technique. (Fig. 3a) was taken with a V30/30P, germanium power transistor. The higher curve was taken with the base open-circuited, giving the I_{co}' curve. The lower curve shows the effect of applying +1·5V reverse bias to the base, to simulate the conditions in a switching circuit. Fig. 3(b) shows the characteristics of a Zener diode, type 652, Fig. 3(c) is the upper trace of Fig. 3(a) with an expanded voltage axis.

The circuit was originally designed to investigate the collector-emitter voltage-current characteristics of high power transistors and the meter ranges were selected for this application. Subsequent tests have shown, however, that the collector-emitter and collector-base characteristics of most transistors, including low power types, may be satisfactorily displayed on an oscilloscope.

Acknowledgments

The author wishes to thank the English Electric Co. Ltd for permission to publish this article and acknowledges the assistance given by Mr. G. M. Ettinger and Mr. R. E. Ireland.

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Two Automatic Impedance Plotters

By R. S. Cole*, B.Sc., Grad.Inst.P., and W. N. Honeyman*, B.Sc.

This article describes two methods by which impedances at microwave frequencies can be automatically presented. It indicates the theory of the two systems and gives the practical details of each together with a comparison of the two.

THE impedance, or Smith[1], chart has proved invaluable in microwave design work but, using the conventional method of plotting, it has the disadvantage of being very tedious and time consuming. This is particularly so when checking a microwave device over a large frequency band or when testing production quantities of microwave components. Using normal methods of plotting an impedance chart it is very easy to overlook sharp resonances and transient effects cannot be observed.

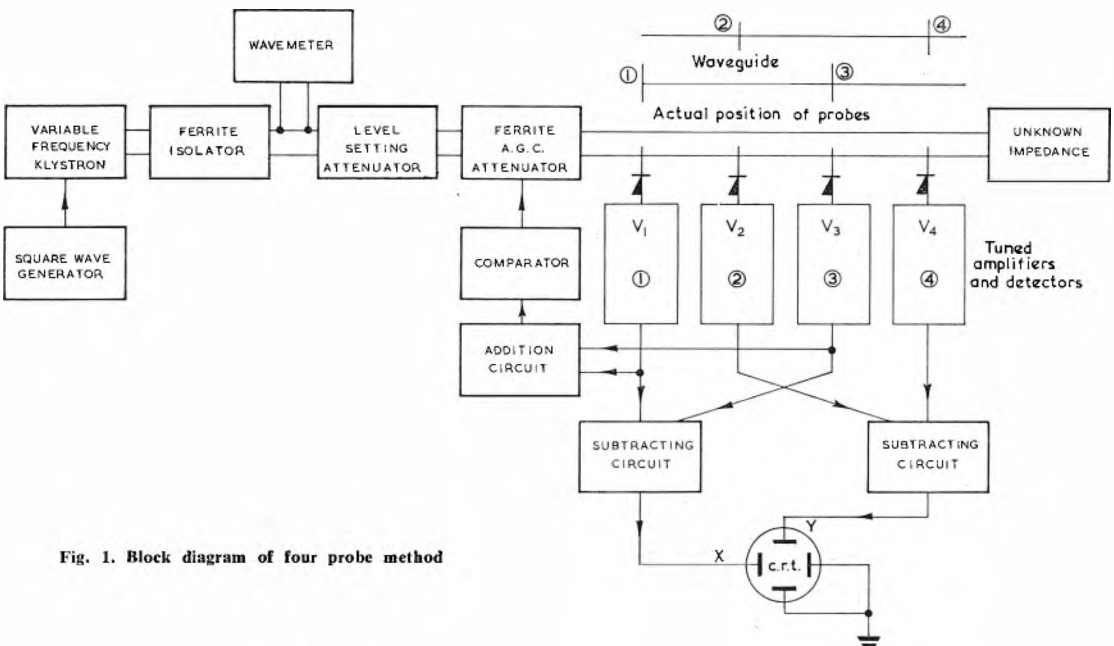


Fig. 1. Block diagram of four probe method

This article describes two types of automatic impedance plotter which have been constructed to display impedance on a c.r.t. Both allow the continuous observation of impedance over a bandwidth of 10 per cent at X-band with an accuracy better than 0·03 in the v.s.w.r. A comparison between the two systems described is given and the inherent errors indicated. These equipments are simple and inexpensive to build compared with other methods of automatic impedance plotting which have been described in the literature[2].

Four-Probe Automatic Standing Wave Gear

THEORY

The property of the Smith chart which is utilized by this method[5] is that any point A on a Smith chart can be represented as a vector $OA = \rho = |\rho|\, e^{j\theta}$, where ρ is the reflection coefficient given by V_r/V_i where V_r is the reflected voltage and V_i is the incident voltage. O is the centre of the impedance chart and θ is the phase-angle between V_i and V_r.

Four probes are spaced in a waveguide at distances of an eighth of a guide wavelength (Fig. 1) and the outputs from the probes are rectified by crystals. When the rectified voltages V_1—V_3 and V_2—V_4 are applied to the X and Y plates of a cathode-ray tube then the impedance of the mismatch under consideration is shown by the trace. The variation of impedance over a frequency band can be shown directly if the frequency is swept. The theory assumes:

(1) The four crystals have a square law.

(2) The amplitude of the microwave power does not vary with frequency.

The incident voltage of the wave at crystal (1) may be expressed as $V_i \cos \omega t$ and the reflected voltage as $V_r \cos(\omega t + \theta)$. With assumption (1) the output from crystal (1) is $[V_i \cos \omega t + \rho V_i \cos(\omega t \mp \theta)]^2$

Therefore $V_1 = (V_i^2/2)\ (1 \mp \rho^2 + 2\rho\cos\theta \mp \text{r.f. terms})$. Crystal (3) is spaced one quarter wavelength from (1) and this gives $V_3 = (V_i^2/2)\ (1 + \rho^2 - 2\rho\cos\theta + \text{r.f. terms})$.

The r.f. terms can be eliminated in detection and so

$$V_1 - V_3 = 2\rho V_i^2 \cos\theta$$

similarly

$$V_2 - V_4 = 2\rho V_i^2 \sin\theta$$

and

$$V_1 + V_3 = V_i^2(1 + \rho^2) = V_2 \mp V_4$$

If $V_1 - V_3$ and $V_2 - V_4$ are applied to the X and Y plates of a cathode-ray tube and V_i is held constant the spot position is given by $X = K\rho\sin\theta$ and $Y = K\rho\cos\theta$ and hence a Smith chart plot is obtained. The factor K takes into account the conversion efficiency of the crystals, the gain of the amplifiers and the deflexion sensitivity of the plates of the c.r.t. The last two of these can be compen-

* E.M.I. Electronics Ltd.

sated by potentiometers in the amplifiers but the conversion efficiency of the crystal is a variable quantity depending on the microwave frequency and other parameters. The selection of four similar crystals becomes important and constitutes the main drawback of this system.

Design

The complete instrument is shown in Fig. 1 and Fig. 2. A klystron is swept continuously over 10 per cent of X-band and is square wave amplitude modulated at 3kc/s. Each of the four crystal outputs is amplified separately by 3kc/s selective amplifiers. The resulting a.c. signals are rectified and the d.c. voltages subtracted in the manner indicated and then applied to the plates of the c.r.t. A ferrite automatic gain control is provided before the crystal holders, the function of which is to keep the term $V_i^2 (1 + \rho^2)$ constant.

The major items of Fig. 1 will be considered separately.

Swept Klystron

The klystron used is the E.M.I. R.5222 in a standard E.M.I. cavity which is tuned by means of two arms normally adjustable by a micrometer movement. The micrometer movement is replaced by a cam designed to give a linear sweep of frequency for 90° rotation. For such an arrangement the value of the reflector voltage for maximum output varies linearly with frequency, and therefore an ordinary linear potentiometer geared to the cam is suitable to sweep the reflector voltage and to keep the klystron oscillating in the same mode throughout the frequency range of the instrument. The cam is driven by a reversible motor which is geared to give a complete sweep of frequency in about ten seconds, a microswitch being used to reverse the motor. It is not convenient to continuously rotate the cam because the requirement for a linear sweep of frequency would make the cam shape unnecessarily complex. Complete rotation of the cam would give four complete cycles of frequency and therefore requires a special potentiometer which would have to be tapped with the appropriate voltages at four points.

Four Probe Crystal Holder

This component is most important in this type of instrument since, with the waveguide matched, the output of the four detector units should be the same over the frequency band of the instrument, and hence the four crystal holders should be identical. With a bad mismatch there is considerable variation of the power incident upon the different crystal units: this provides a serious limitation with bad mismatches since the detector law of the crystals varies considerably with the incident power. By the use of selected crystals and with the a.g.c. device in operation matches down to 0·6 could be measured fairly accurately.

Due to the physical size of the four crystal holders they could not be separated by an eighth of a wavelength by spacing on one side of the waveguide in an X-band system. It was more convenient to have two holders on each side of the guide spaced appropriately (Fig. 1). The spacing of the probes, however, is only correct at a single frequency and at any other frequency the resulting trace will not represent exactly the impedance in any plane. For a true impedance plot it is necessary to alter the probe spacing with frequency and then the impedance plot refers to the fixed plane with reference to which all the probes are moved. This plane could be one of the probes or the central plane of the probes (which entails minimum mechanical movement of any probe) or any plane along the waveguide chosen at will.

If the probes are fixed the trace will most accurately represent the impedance in the plane which would require the minimum of probe movement, i.e. the central plane of the four probes. The errors involved in this assumption can be evaluated:

If voltages at the centre of the probe system are $V_i \cos \omega t$ and $V_R \cos (\omega t + \theta)$, then the voltages at probe (1) are $V_i \cos (\omega t - 3\delta\pi/8)$ and $V_R \cos (\omega t + \theta + 3\delta\pi/8)$, where $\delta = \lambda/\lambda_0$, λ is the wavelength being considered and λ_0 is the design wavelength.

Similar expressions apply for the voltages at the other probes. The output from each mixer can be obtained assuming square law detection and the position of the c.r.t. trace will be given by:

$$V_x = V_{(1)} - V_{(3)} = -2V_i V_R \sin (\theta + \delta\pi/4) \sin \delta\pi/2$$

$$V_y = V_{(2)} - V_{(4)} = -2V_i V_R \sin (\theta - \delta\pi/4) \sin \delta\pi/2$$

The radius of this point is given by:

$$r^2 = V_x^2 + V_y^2 = 4V_i^2 V_R^2 (\sin^2 \delta\pi/2)(1 - \cos 2\theta \cos \delta\pi/2)$$

and the fractional error in radians by:

$$(1 - \cos 2\theta \cos \delta\pi/2)^{\frac{1}{2}} \sin \delta\pi/2 - 1$$

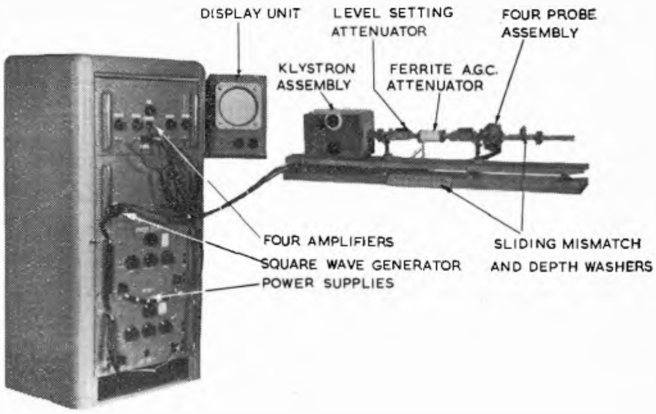


Fig. 2. Four probe system

The radius is at an angle $\varphi = \tan^{-1} (V_x/V_y)$ and so

$$\tan \varphi = \frac{\sin (\theta + \delta\pi/4)}{\sin (\theta - \delta\pi/4)}$$

The angular error (in radians) is thus:

$$\tan^{-1} \left(\frac{\sin (\theta + \delta\pi/4)}{\sin (\theta - \delta\pi/4)} \right) + \theta + \pi/4.$$

The maximum errors introduced by this system occur at the ends of the 10 per cent bandwidth and are ± 4 per cent in the reflection coefficient and $\pm 2°$ in the phase angle.

Ferrite A.G.C. System and Ferrite Attenuator

Since this instrument depends upon the incident voltage V_i being constant with frequency an automatic gain control system is essential.

The method finally adopted uses a variable ferrite attenuator. The sum of the voltages V_1 and V_3, i.e. $V_i^2 (1 + \rho^2)$ is compared to a standard reference voltage and the difference used to control the current through the ferrite attenuator, Fig. 3. Since this method holds $V_i^2 (1 + \rho^2)$ constant and not V_i^2 a small error is involved. However, with a v.s.w.r. of 0·6 this error is only 6 per cent in reflection coefficient and diminishes as the match improves: it can be completely eliminated by adjusting the Smith chart cursor.

An alternative method is to use the difference between $(V_1 + V_3)$ and the standard reference voltage as the bias for variable-mu valves in the four identical amplifiers. This requires that the control characteristics of the four variable-mu valves should be identical and, in practice, would mean selection of valves.

The ferrite attenuator[7] relies for its operation on Faraday rotation of the energy into a piece of resistive card placed alongside the ferrite. A rotation of the electric field vector over 90° results in absorption in the card as it rotates through 90° giving rise to a broad-band matched attenuator. The attenuation depends on the degree of rotation which in turn depends upon the energizing current.

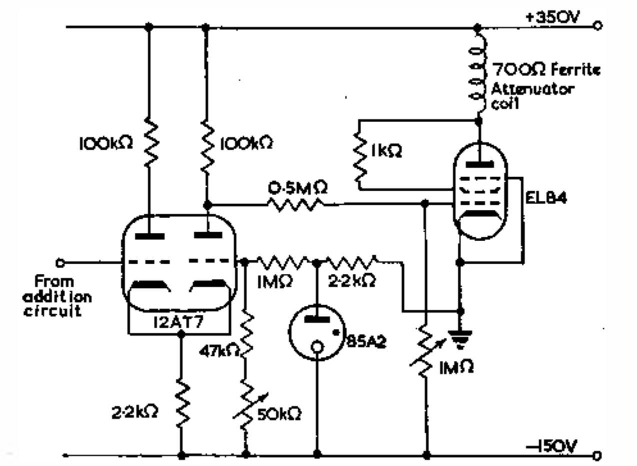
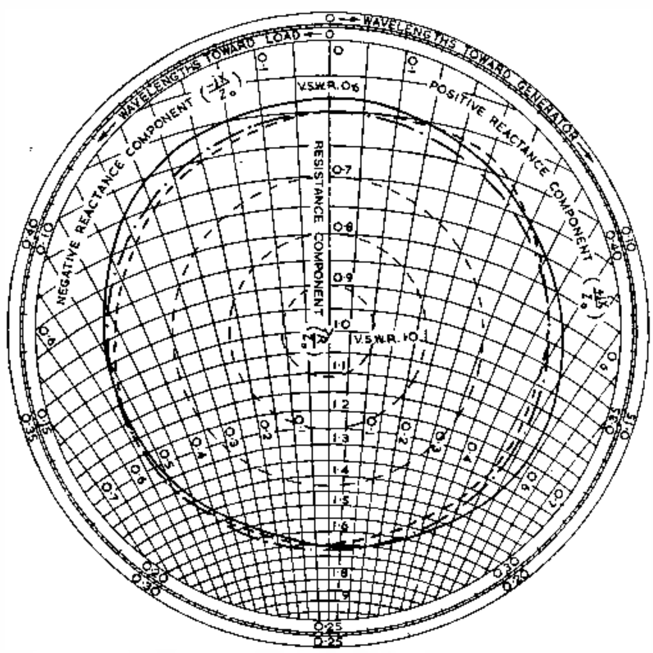


Fig. 3. Comparator circuit for automatic gain control



———— f_0 +5% — · — f_0 - - - - - f_0 − 5%

Fig. 4. Impedance chart of four probe method

Amplifier and Display System

One separate linear amplifier comprising three EF86 valves and twin-T selective filter networks tuned to 3kc/s is used to amplify the output of each crystal. To obtain a deflexion proportional to $V_1 - V_3$ or $V_2 - V_4$ the rectified outputs of the two corresponding amplifiers are applied to the grids of a long tailed pair and the plates of the c.r.t. are connected to the anodes. The amplifiers are provided with gain controls which can be adjusted to compensate for any small differences in the four signal channels.

An alternative system subtracts the outputs of the crystals before amplification but the sign of the resultant voltages is lost unless phase sensitive amplifiers are employed.

Performance

This was checked by means of a slotted line and a sliding probe, the depth of which could be varied and hence any v.s.w.r. at any phase could be simulated. Fig. 4 shows the trace of the impedance of a matched load over the frequency range of the instrument. The v.s.w.r. over the centre 5 per cent of the band is better than 0·98 and over the whole 10 per cent band better than 0·97. Also on the same figure the trace at all phases of 0·6 v.s.w.r. mismatch is shown at the extreme ends and the centre of the 10 per cent frequency band. At the centre the error is negligible while at the ends of the band the trace is elliptical and also slightly displaced. However, total errors do not exceed ±0·03 on the v.s.w.r. Phase errors are not greater than those calculated theoretically.

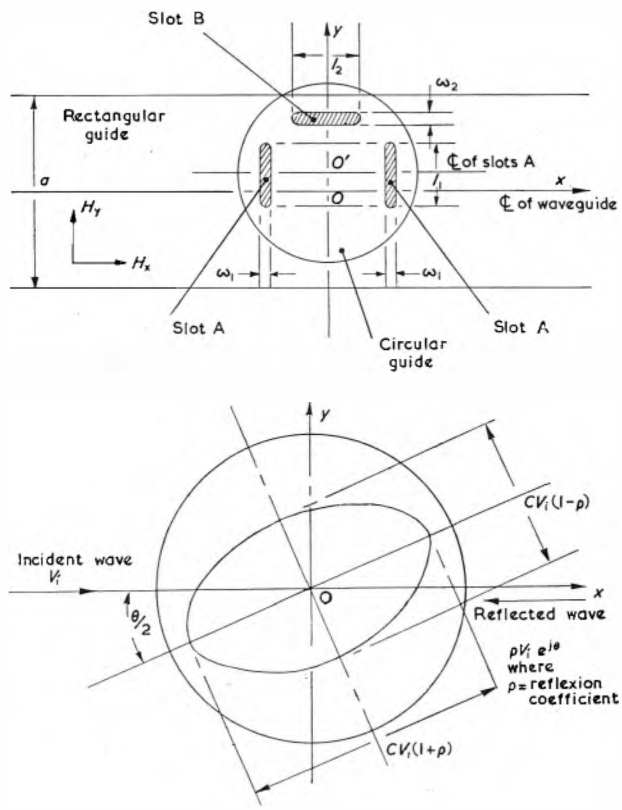


Fig. 5. (a) Three slot, rectangular-circular coupler
(b) Field pattern excited in circular guide

Rotating Crystal Method

This method of automatic impedance presentation depends upon the properties of a three slot waveguide coupler from rectangular to circular guide described by S. B. Cohn[3,4]. This coupler consists of a circular waveguide joined to the broad face of a section of rectangular guide, magnetic coupling taking place through three slots as shown diagrammatically in Fig 5(a). In this figure, the two slots A couple the H_y magnetic field and the single slot B couples the H_x field between the rectangular to circular guide. A consideration of the phase differences normally existing in rectangular guide field patterns shows that an H_{10} mode travelling in the direction x in the rectangular guide sets up a circularly polarized H_{11} mode in the circular guide. Similarly, a wave travelling in the opposite direction sets up a second circularly polarized wave, but with the opposite hand of polarization. Incident and reflected waves exist simultaneously and the two circularly polarized waves set up in the circular guide combine to give an elliptically polarized wave as shown in Fig. 5(b). The ratio of the minor to major axes of this ellipse can be shown to be equal to the standing wave ratio in the rectangular guide and the angle between the major and x axis shown to be half the phase angle of the reflected wave (Fig. 5(b)). Hence the field pattern in the circular guide contains all the information necessary for an impedance presentation and since the coupler can be made broad-banded it forms the basis of an automatic impedance plotter.

the axis of the circular guide at 50c/s. Hence the waveform out of the crystal holder, assuming a square law crystal, consists of a square wave, modulated by a sinusoidal waveform, the modulating frequency being twice the crystal holder rotational frequency. The minimum and maximum amplitudes of the output signal are proportional to the square of the minor and major axes of the elliptical field pattern respectively. Using the notation given in Fig. 6, the amplitude of the modulation envelope is 4 $KV_i^2\rho$ and the mean level is KV_i^2 $(1 + \rho^2)$ where K is a constant depending upon the conversion efficiency of the crystal. This signal then passes into an a.g.c. amplifier, which holds the mean

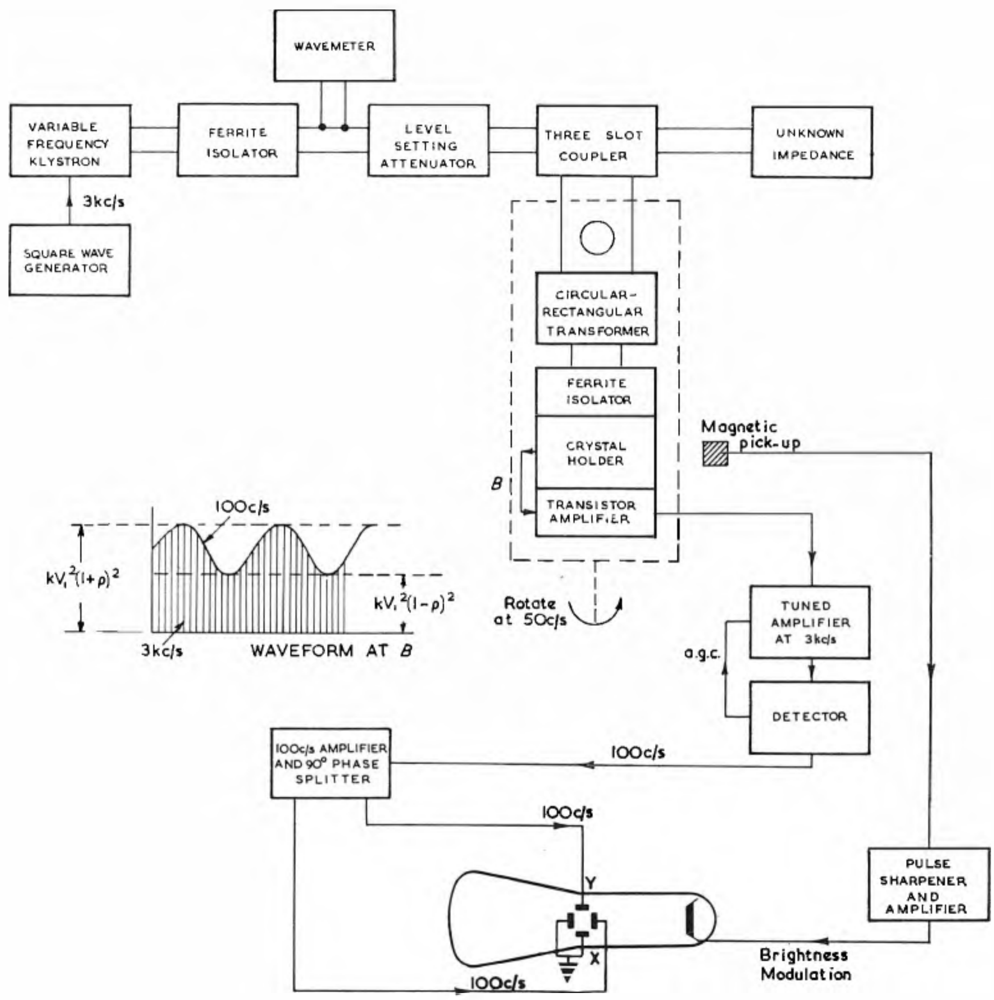


Fig. 6. Block diagram of rotating crystal method

THEORY OF THE ROTATING CRYSTAL SYSTEM

The method of obtaining the information from the field patterns in the circular guide is illustrated in Fig. 6, showing a block diagram of the apparatus. The principle of this method was demonstrated by the British Thomson-Houston Co. Ltd at the 1956 Physical Society Exhibition.

The microwave signal is amplitude modulated by a 3kc/s square wave and passes via a ferrite isolator and level setting attenuator into the three slot coupler and then into the unknown impedance which sets up its characteristic incident and reflected waves. These set up the elliptical field patterns as described above, which travel to a circular-rectangular transformer. The transformer transmits one direction of electric field only which is detected by a crystal holder; the field at right-angles being absorbed in a resistive card. The crystal holder and transformer are rotated about

level of the output constant, so that the resulting signal out of the detector is proportional to 4 $\rho/(1 + \rho^2)$. The remaining circuits produce a circle on a c.r.t. whose radius is proportional to this detected signal. A true representation of a Smith chart would require a circle of radius ρ, but the inaccuracy this introduces is negligible for standing wave ratios greater than 0·6 and for worse standing wave ratios a special cursor was constructed to eliminate these errors.

To give a complete impedance chart representation the phase angle must be added and this takes the form of a pulse which is obtained once per revolution at a particular position of the rotating crystal holder. Since the angle between the major axis of the electric field ellipse and the x axis is half the phase angle of the reflected signal from the unknown impedance, and the modulating frequency is twice the frequency of rotation of the crystal holder, a true impedance chart is produced if this pulse is used to brightness modulate the c.r.t.

DESIGN OF THE ROTATING CRYSTAL IMPEDANCE PLOTTER

The major items will be considered separately.

KLYSTRON AND RECTANGULAR GUIDE COMPONENTS

The klystron used is the same unit as described in the four probe method. The field displacement ferrite isolator is inserted to prevent mismatches pulling the klystron, the isolator having 15dB backward loss over the 10 per cent frequency band. A level setting attenuator is used to prevent the amplifiers overloading from excessive signals.

THREE SLOT COUPLER

Since the performance of the impedance plotter depends mainly on this coupler, its design is of great importance. To obtain a final design, one of the dimensions in the coupler was varied until the performance was an optimum. To obtain a wide-band design it was found that the width of the slots must be kept as small as possible, probably to avoid coupling magnetic fields at right angles to the slot. Off-setting the centre of the circular guide from the centre of the rectangular guide also improved the bandwidth probably because the H_y field decreases as it approaches the side wall, thus reducing the coupling of this component through slot *B*. The easiest method of producing these slots accurately is by using spark erosion, since a repeatability of ·0.003in can be achieved. The coupler finally adopted has an axial ratio better than 0·98 and its coupling is approximately constant at 35dB over the 10 per cent band.

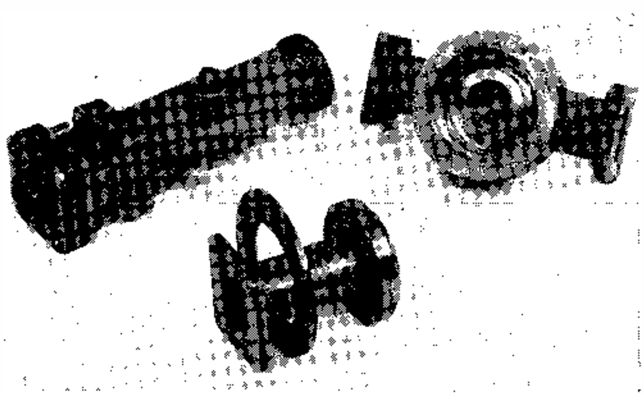


Fig. 7. Rotating crystal system

ROTATING COMPONENTS

A standard choked rotatable section in circular guide is used to separate the moving and stationary components as shown in Fig. 7. Also shown is the stepped circular-rectangular transition which uses resonant iris matching. A piece of resistive card suitably tapered is placed at right-angles to the direction of propagation to absorb any electric field not transmitted.

Because the signal output originates from the crystal holder which is rotating at 3 000rev/min, one of the basic problems in this instrument is the connexion of the signal from the rotating system to the amplifier. Since a klystron is used as a microwave source and the three slot coupler gives 35dB insertion loss, the crystal holder signal is only of the order of 10mV. A mercury contact was used initially in which two steel cylinders rotated in mercury baths which were almost sealed from the atmosphere. This type of device was satisfactory for short periods but after long running times, dirt found its way into the system resulting in low signal-to-noise ratios and consequent unsteady traces. A potted transistor amplifier, having a gain of about 100 is now used between the crystal and the output and ordinary rhodium plated contacts with brushes are used with the relatively large signal so produced. Variations in the gain of this amplifier do not affect the system since the external amplifier is a.g.c. controlled; the only requirement of the amplifier is linearity.

The whole system is rotated by a synchronous $\frac{1}{8}$ h.p. motor driving in line with the axis of the circular waveguide.

MAIN AMPLIFIER

The design of this amplifier and detector unit is conventional since the input signal is large. The a.g.c. loop requirements are stringent so that both backward and forward acting a.g.c. bias on four variable-mu valves is included. This bias is generated by a comparison between a reference voltage and the mean detected voltage so that a relatively stable, flat a.g.c. characteristic is obtained over a 10:1 variation of input signal. By varying this reference voltage the inner portion only of the Smith chart can be used, thus giving greater accuracy.

PULSE BRIGHTENING SYSTEM

The pulse necessary for this system is derived from a tape recorder eraser head when a magnet attached to the crystal holder passes it. This system gives a pulse of about one volt and after squaring, differentiating, clipping and amplifying the resulting signal is 0·1msec long and 100 volts in amplitude.

PERFORMANCE

The instrument was checked by the same method as used in the four probe case. With a matched load a v.s.w.r. of not less than 0·98 over the 10 per cent frequency band was recorded. Using a 0·6 v.s.w.r. sliding mismatch errors did not exceed $\pm 0{\cdot}02$ on the v.s.w.r.

Conclusion

The main disadvantage of the four probe method is the selection of four similar crystals and the need to ensure identical conditions in the four channels. When a perfect load is measured the two sets of amplifiers are balanced to give a spot at the centre of the Smith chart and accuracy over a period requires that there should be negligible amplifier drift. However, this instrument requires no moving parts which are necessary on the rotating crystal type, but the latter has an advantage in that it uses only one crystal. Inherently the rotating crystal method has fewer inaccuracies which is shown in a comparison of the performances.

Both instruments are suitable for rapidly carrying out the great majority of production testing of microwave components and systems and for curtailing development work of an empirical nature.

Acknowledgments

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Some Aspects of the Performance of

TELEVISION MAINS-HOLD CIRCUITS

By R. D. A. Maurice*, Ing. Dr., Ing. E.S.E., A.M.I.E.E.

Phase instability of the electricity supply to a television studio or outside broadcast centre causes transient changes in the repetition frequency of field-synchronizing pulses emerging from the waveform generator. These transients cause the received television image to bounce vertically, particularly during transmission of film by continuous motion scanning or during film recording. The effect of film-traction mechanical filters upon these transients is examined and recommendations are made regarding methods of specifying mains-hold performance so as to avoid excessive picture bounce.

THE waveform generator is that part of a television channel devoted to the generation of signals required for the control of timing, and hence picture geometry as distinct from the picture-modulation signals derived from scanning an image of the scene to be transmitted. Among the signals produced by the waveform generator are the line and field synchronizing pulses, and it is customary[1] to ensure synchronism between the fundamental component of the (repeated) field-synchronizing pulses and the frequency of the supply mains. The reason for this is that any residual hum either in receivers or transmission equipment which gives rise to light and dark regions of the picture and slight distortions of the raster will be stationary and thus much less evident to the viewer.

A difficulty arises, however, in that although the supply-mains frequency is practically constant†, its phase is subjected to sudden transients due to the switching in and out of circuit of neighbouring loads. The mains circuits, although of relatively low impedance, cannot be regarded as being possessed of perfect regulation and thus, while a power station's rotating machine shaft may be regarded as a source of constant-frequency power devoid of transients, this is far from true of the user's terminals. The problem of synchronizing television-field pulses with the frequency of the supply mains is thus a real one. The generally accepted solution at present is to incorporate in the waveform generator a circuit called a 'mains-hold'. This may consist[1] of a rotary master-frequency generator driven from the mains supply by a synchronous motor through a mechanical low-pass filter. Modern technique seems to favour electronic circuits using a phase discriminator and a reactance-valve controlled oscillator in a feedback loop.

Whatever method is used, it remains true that a transient in the phase of the mains supply must give rise to a transient change in the repetition frequency of the field-synchronizing waveform in all television systems which are locked to the mains, however loose the lock, unless the reference mains waveform be obtained from an undisturbed, transient-free source.

A sudden change in synchronizing-pulse repetition frequency will cause no objectionable effect in a television receiver if the time-base control circuits have short time-constants enabling the scanning waveforms to adapt themselves to the new conditions in less time than one scanning period. As regards changes of field-synchronizing-pulse repetition frequency, the above conditions apply even in receivers using 'flywheel time-bases' as the 'flywheel' effect applies to the line time-base and not to the field. A horizontal 'bounce' may, however, result from the same cause, since the line and field repetition frequencies are essentially locked.

The purpose of this article is to examine and clarify the effects caused by mains-phase transients after passage through mains-hold circuits when the television equipment in service includes circuits or devices unable to respond with adequate rapidity. In particular, what follows will apply directly to continuous motion film scanners and television film-recording machines, because they must contain mechanical low-pass filters in order to ensure the greatest possible uniformity of film velocity to avoid sprocket-hole flutter, 'wow', and gear or drive velocity fluctuations.

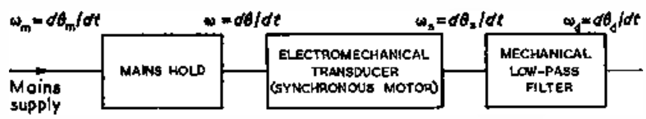


Fig. 1. Fundamental circuit

The Fundamental Problem

The schematic diagram (Fig. 1) shows the fundamental circuit through which a mains phase transient must pass, using a continuous-motion television recording system as the vehicle for investigation.

The mains-hold supplies a waveform whose fundamental frequency is 'locked' to that of the supply mains by a relationship which will be dealt with later. This waveform is fed to an ideal synchronous motor in which the angular velocity of the shaft is rigorously proportional to the frequency of the mains-hold output. The motor drives the sprocket wheel of the film-drive mechanism and the latter is arranged to form a low-pass filter in such a way that the film is drawn, by the sprocket wheel, over a smooth drum whose angular velocity can contain only components whose frequencies are within the filter pass-band. A simple example, which will be used frequently in this article, is that of the Davis filter[2] shown in Fig. 2, reproduced from the Journal of the Society of Motion Picture Engineers.

The mains frequency is $\omega_m/2\pi$, a function of time. The frequency of the fundamental component of the mains-hold output of field-pulse waveform will be denoted $\omega/2\pi$. If q is a factor representing the number of poles on the motor and also any gear ratios between motor shaft and film-drive sprocket wheel, the angular velocity of the latter may be denoted $\omega_s = \omega/q$. The angular velocity of the film drum or mechanical filter output, ω_d, will be related to the sprocket-wheel or mechanical-filter input angular velocity by the complex transfer function of this filter G [jα]; where $\alpha/2\pi$ is the 'frequency' of variation of the fundamental frequency of mains-hold output waveform.

The concept of 'transfer function', so familiar to those

* *The British Broadcasting Corporation.*

† *Rates of change of British Grid-System frequency of the order of 0·001c/s/s have been observed: thus it would take 17m. to change frequency from 50c/s to 51c/s, for example. ("Muirhead Technique", Vol. 6, No. 4.)*

versed in the art of electric wave filters and electrical network theory in general, may be applied to linear dynamic systems with great profit; thus, if one has a rotating mechanical system in which an input or driven shaft is coupled by any linear combination of mass, stiffness and viscous friction (friction in which the force is proportional to velocity) to an output or driving shaft, one finds that the dynamic equation obtained by applying d'Alembert's principle* is analogous to a suitably chosen electrodynamic equation in which inductance is the analogue of moment of inertia, capacitance is the analogue of compliance (reciprocal stiffness), resistance that of viscous friction,

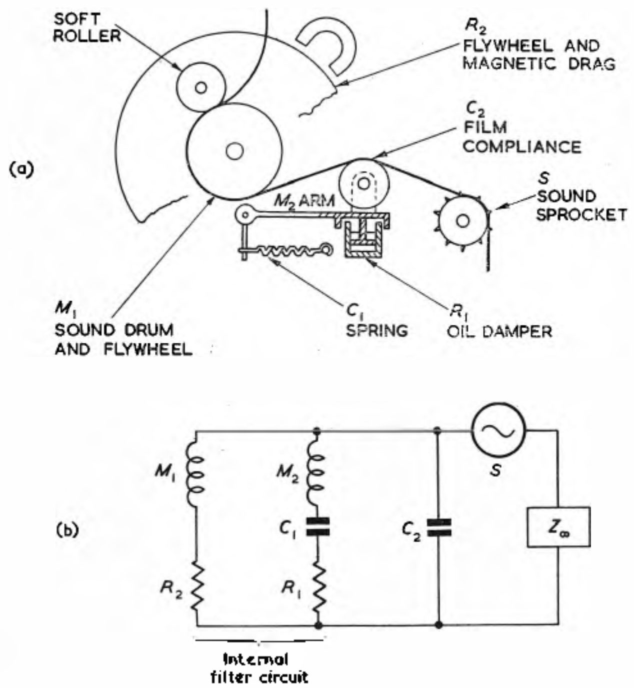


Fig. 2. Single arm film-drive mechanical filter (Davis filter)
(a) The mechanical filter
(b) Equivalent electric circuit

electric current of angular velocity, electric charge of angle (of rotation), etc. It can be seen, therefore, that just as one may consider the complex current-transfer-function of a linear four-terminal network as expressing the variation with frequency of the ratio of the output current to the input current, so by analogy one may consider the complex angular-velocity transfer-function of a linear rotating dynamic system as expressing the variation with frequency (of the angular velocity) of the ratio of the output (shaft) angular velocity to that of the input (shaft).

When the mains frequency $\omega_m/2\pi$ is constant, the mains-hold suffers no disturbance and may be expected to supply a constant-frequency signal to the synchronous motor, which in turn will feed the mechanical filter with a constant angular velocity. The filter being a low-pass device, that is, able to pass a constant angular velocity from input to output without attenuation, will communicate this angular velocity to the smooth drum and one may write $\omega_s = \omega_d =$ constant. Now the film displacement or position in the gate will be directly proportional to the angle of the drum, θ_d, while the position, in the direction of film movement, of the flying spot or light beam representing picture signals will be proportional to the angle of the synchronous-motor shaft, θ_s. Assuming that at a certain time arbitrarily chosen as $t = 0$ a film frame is correctly located in the gate and that the recording light beam is likewise correctly framed in the same gate, then $\theta_s - \theta_d = 0$ and all further television images will be perfectly framed in their respective film frames because all velocities have been assumed to be constant and the initial condition of perfect framing must thus be maintained.

If, now, the mains frequency changes, the mains-hold output phase must depart from linearity with time and the shaft of the synchronous motor must change its angular velocity; ω_s will give rise to a change in ω_d and thus both the television images (or light beam) and the film frames will alter their positions (or timings) in the gate. It might be supposed that if the mains-hold output changed its frequency slowly enough, that is, was sinusoidally frequency modulated by a low enough modulation frequency, there would be no mis-alignment between the television image and film frame in the gate because the mechanical filter constituted by the film transport mechanism would be able to follow faithfully such slow variations in frequency. This is not so, however, because mis-alignment or framing error between the television image and film frame is proportional to a difference between two phases or angles which are themselves the integrals of the frequency changes involved. The mechanical filter will have a delay which, although becoming increasingly constant for very low-frequency changes in angular velocity, may none the less tend to a finite time delay for zero frequency change, and so the difference between the mechanical filter input and output shaft phases or angles becomes the difference between the integrals of two identical functions of time separated by a constant interval. Such a difference is not normally zero.

In practice, if the mains frequency changes very slowly in a given direction and is followed perfectly by the mains-hold, the result is a gradual shift of framing or positioning of the recorded television images vis-à-vis the film frames containing them. If the frequency change were sufficient it would be necessary to readjust the framing of the film so as to compensate for this error. Fortunately the supply mains change frequency slowly and the problem does not present itself in this form during the recording of a single television programme if an appropriate mechanical filter is used.

It was mentioned earlier that the main source of trouble is switching-transients, which cause sudden changes in mains frequency. Such a change may more conveniently be regarded as the addition of a unit step change of phase to the instantaneous phase of the mains at any given instant, say at $t = 0$. From the foregoing explanation it will be evident that such a transient will cause a more or less rapid departure from correct alignment of television image and film frame in the film gate. This variable framing error is often referred to as picture bounce, and, within limits, is tolerable if not too great in amount or too rapid in effect.

One object of a mains-hold performance specification should be to restrict the amount and rapidity of picture bounce when television equipment is being used which contains systems having considerable inertia. It will be apparent that a precise specification is only possible when the exact form of the mechanical filter in use as well as the detailed behaviour of the mains-hold are known.

In what follows, some specific examples will be examined with a view to showing the methods to be adopted in any particular case. First the relationships between frequency and phase which will form the basis of the remaining analysis, will be formulated.

* *That in a linear dynamic system there is, at any instant of time, equilibrium between the forces of inertia, the forces of constraint and the external applied forces.*

The mains-hold output at field-pulse repetition frequency is a voltage cos θ where in the absence of disturbances the phase θ is a linear function of time

$$\theta = \omega_0 t \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

The frequency of this output may be found from the formula

$$\omega = d\theta/dt$$
$$= \omega_0 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

In the United Kingdom one may take $\omega_0/2\pi$ as 50c/s.

After a time, t, the mechanical filter input shaft (which may also be the ideal synchronous motor shaft) will have turned through an angle θ_s radians where

$$\theta_s = \int_0^t \omega_s dt = 1/q \int_0^t \omega dt \quad \ldots\ldots\ldots \quad (3)$$

During this same time interval the mechanical filter output drum will have turned through the angle θ_d where

$$\theta_d = \int_0^t \omega_d dt \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

It will be seen later that, in general, $\theta_d \neq \theta_s$. There will thus arise a framing error. Suppose that N is the number of television fields which will have been presented to the film gate in the time t and that ΔN is the difference between this number and the correct number if perfect alignment between television-field presentation and film-frame positioning were to be obtained. Evidently

$$\Delta N/N = (\theta_s - \theta_d)/\theta_s \quad \ldots\ldots\ldots\ldots \quad (5)$$

But television fields will be presented to the film gate in accordance with mains-hold output at field frequency, thus

$$N = (1/2\pi)\int_0^t \omega dt \quad \ldots\ldots\ldots\ldots\ldots \quad (6)$$

From equations (6) and (3)

$$N = q\theta_s/2\pi \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

and from equations (7) and (5)

$$\Delta N = (q/2\pi)\ (\theta_s - \theta_d) \quad \ldots\ldots\ldots\ldots \quad (8)$$

This equation relates the framing error to the angle difference between mechanical-filter input and output.

Transient Variation in Mains-hold Field-pulse Repetition Frequency

This condition applies to all cases in which the mains-hold output waveform, whether due to a change of phase of mains voltage or not, departs only temporarily from a uniform condition implying a fundamental component of constant frequency.

The actual analysis required must depend upon the exact form taken by the frequency of the fundamental component of the mains-hold output during the transient condition set up by, say, an abrupt change of phase in the mains supply.

In general terms the problem can be stated as follows: let the mains-hold output waveform have a fundamental component of frequency $\omega_0/2\pi$ except during a transient condition due perhaps to a unit-step phase change in the mains supply voltage. During this transient period let the mains-hold output have a fundamental component of frequency $[\omega(t) + \omega_0]/2\pi$. If this waveform be fed to a synchronous motor, defined for this purpose as capable of faithfully following mains frequency variations, then the motor shaft will rotate with an angular velocity proportional to $\omega_0 + \omega(t)$ and the sprocket wheel drawing the film through the film traction mechanism will have an angular velocity

$$\omega_s(t) = 1/q\ [\omega(t) + \omega_0]$$

Let the mechanical filter output drum, over which the film passes on its way through the gate, rotate with an angular velocity $\omega_d(t)$.

Square brackets, [ ], will be used to indicate the 'operational equivalent' of a function of time written in parentheses, ( ); thus

$$\omega_s[p] \simeq \omega_s(t)$$
$$\omega_d[p] \simeq \omega_d(t)$$

where p is the operator d/dt and equals jα* for a steady-state regime at frequency $\alpha/2\pi$.

Let the transfer function; 'output-angular-velocity against input-angular-velocity': of the mechanical filter be $G[p] = G[j\alpha]$†, whence

$$\omega_d[p] = \omega_s[p]\ .\ G[p] \quad \ldots\ldots\ldots\ldots \quad (9)$$

The operational way of writing equations (3) and (4) is

$$\theta_s[p] = (1/p)\,\omega_s\,[p] \quad \ldots\ldots\ldots\ldots \quad (10)$$
$$\theta_d[p] = (1/p)\,\omega_d\,[p] \quad \ldots\ldots\ldots\ldots \quad (11)$$

whence

$$\theta_s(t) - \theta_d(t) \simeq \theta_s[p] - \theta_d[p] = 1/p(\omega_s[p] - \omega_d[p]) \quad \ldots\ldots \quad (12)$$

Substituting equation (9) into equation (12) and the result into equation (8) and replacing ω_s by ω/q

$$\Delta N \simeq 1/2\pi\ .\ \omega[p]/p\,(1 - G[p]) \quad \ldots\ldots \quad (13)$$

or, if preferred,

$$\Delta N \simeq 1/2\pi\ \theta[p]\,(1 - G[p])$$

Thus, given the filter transfer function, $G[p]$ and the transient waveform of the mains-hold output, $\omega(t)$, due to a disturbance at its input, such as an abrupt phase change in the mains; one may obtain $\omega[p] \simeq \omega(t)$ by operational (or classic) methods, substitute it into equation (13) and find ΔN as a function of time, again by operational or classic methods. One then has to find the maximum value ΔN_1 of ΔN, and if one gives it a limiting, tolerable value, one may then find the maximum permissible frequency deviation $\Delta\omega/2\pi$. The limiting tolerable value of ΔN_1 which will be used henceforth will be $\frac{1}{4}$ of a television line or $\frac{1}{4} \times 1/405 = 1/1600$‡, approximately. This figure was found experimentally by applying sinusoidal 'field-shift' to a picture tube displaying an unmodulated 405-line raster. A fine grating pattern of horizontal opaque and transparent lines was placed immediately in front of the display-tube face plate. The resulting moiré or 'beat' pattern was caused to change at the frequency of the 'field-shift' waveform. At a 'field-shift' frequency of $\frac{1}{2}$c/s a quarter-line picture bounce was found to be tolerable even under the foregoing stringent conditions.

Having outlined the method, one or two applications will be considered.

The Fundamental Component of the Mains-Hold Output Waveform has a Frequency of which the Rate of Change is Constant

This case, although supplying a necessary condition, is not regarded as sufficient for a mains-hold specification as it does not represent the normal response of a typical mains-hold to a mains-phase transient, nor does it suit certain types of crude mechanical filter.

Assume that the mains-hold output changes frequency at the constant rate of mc/s in a second. Then

$$m = (1/2\pi)\ .\ d\omega/dt \quad \ldots\ldots\ldots\ldots \quad (14)$$

---

* *$\alpha/2\pi$ is the frequency of a steady state sinusoidal angular velocity used as a mechanical filter "test" function in the same way as a sinusoidal voltage may be used to find the frequency/response function of an electric wave filter.*

† *It should be remembered that $G(j\alpha)$ is the transfer function for any common quantity used as input and output, provided these are linearly related to angular velocity; thus, instead of angular velocity, one may use angular (or linear) displacement, or even angular acceleration, if desired.*

‡ *The figure 1/1600 is purely nominal. If the dimensions of standard 35mm film be taken into account, it can be shown that $\Delta N_1 = 1/1600$ corresponds with a picture bounce of approximately 1/7 of a television line.*

If $\omega/2\pi$ is regarded as a transient departure from a constant frequency $\omega_0/(2\pi)$ one may write, from equation (14), neglecting the constant of integration ω_0,

$$\omega(t) = 2\pi mt \quad \ldots\ldots\ldots\ldots\ldots \quad (15)$$

whence from a simple operational rule

$$\omega[p] = 2\pi\, m/p \quad \ldots\ldots\ldots\ldots\ldots \quad (16)$$

Mechanical Filter of Transfer Function $G[p] = \alpha_c/(\alpha_c+p)$

This simple example of a transfer function and equation (16) can now be substituted into equation (13), thus obtaining

$$\Delta N \simeq (m/p^2)\ [1 - \alpha_c/(\alpha_c + p)] \quad \ldots\ldots \quad (17)$$

By operational or classic methods

$$\Delta N = (m/\alpha_c^2)\,[\alpha_c t - \exp(\alpha_c t)] \quad \ldots\ldots\ldots \quad (18)$$

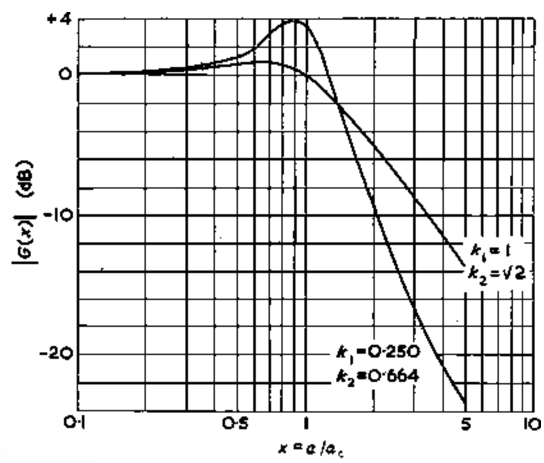


Fig. 3. Transmission/frequency characteristics of Davis filter
(Transmission may be either in terms of angular velocity or in terms of angular displacement)

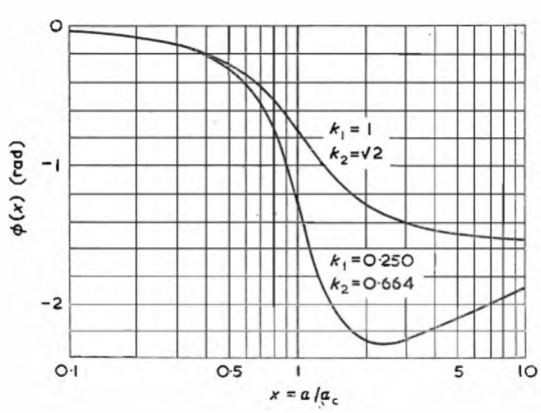


Fig. 4. Phase/frequency characteristics of Davis filter
(Phase may be taken either in terms of angular velocity or in terms of angular displacement)

This function does not present a maximum but continues to increase indefinitely, so that, having a fixed maximum tolerable value $\Delta N_1 = 1/1600$, for ΔN, one may find at what time, t_1, this occurs; but as $\omega(t) = 2\pi\, mt$ the value $\omega_1 = 2\pi\, mt_1$ will represent the maximum permissible departure from the constant or steady value, $\omega_0/(2\pi)$. Fig. 4, curve 1 shows the maximum permissible frequency deviation, $\omega_1/(2\pi)$, as a function of rate of change of frequency, m, for $\alpha_c/(2\pi) = 1$c/s, (a typical value of cut-off frequency for film traction filters). The figure $m =$ 0·075c/s/s corresponds to a rate of change of frequency of 0·15 per cent a second if the steady frequency, $\omega_0/(2\pi)$, is 50c/s. This particular value has at times been used in specifications, but it will be observed that unless accompanied by a limit to the actual frequency deviation, it has little significance with relation to framing error for the type of mechanical filter considered in this section.

Mechanical Filter of Davis Type (Figs. 2, 3 and 4)

Here, neglecting the mass M_2 of the jockey-pulley arm and the film compliance C_2,

$$G[p] = \frac{\alpha_c^2 + k_1\alpha_c p}{(p + \frac{1}{2}k_2\alpha_c)^2 + \alpha_c^2\,(1 - k_2^2/4)} \quad \ldots\ldots\ldots \quad (19)$$

where k_1 and k_2 are dimensionless constants which determine the performance of the filter.

Substituting equations (19) and (16) into (13)

$$\Delta N \simeq (m/p^2)\left[1 - \frac{\alpha_c^2 + k_1\alpha_c p}{(p + \frac{1}{2}k_2\alpha_c)^2 + \alpha_c^2(1 - k_2^2/4)}\right]. \quad (20)$$

whence, more conveniently by operational methods,

$$\Delta N = (m/\alpha_c^2)\left\{1 + k_1k_2 - k_2^2 + \frac{k_2 - k_1}{\sqrt{(1 - k_2^2/4)}}\ zt + \frac{\exp[-zt/\sqrt{(4/k_2^2 - 1)}]}{\sqrt{(1 - k_2^2)]}}[\sin(zt+3\phi) - k_1\sin(zt+2\phi)]\right. \quad \ldots\ldots\ldots\ldots \quad (21)$$

where $z = \alpha_c\sqrt{(1 - k_2^2/4)}$ and $\phi = \arctan\sqrt{(4/k_2^2 - 1)}$.

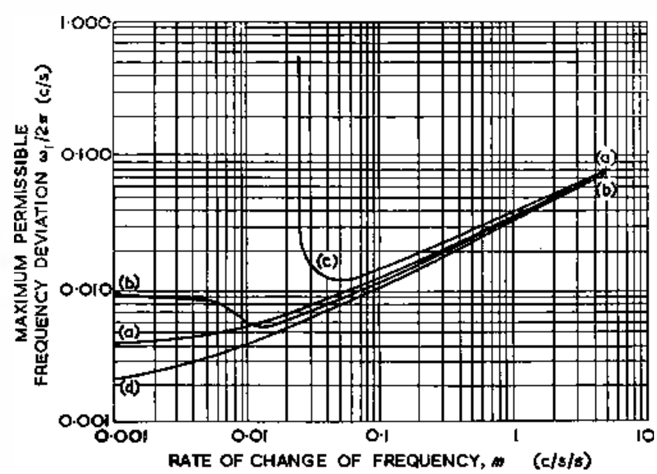


Fig. 5. Transient deviation of field frequency giving rise to a ¼ line picture bounce: field frequency varying with a constant rate of change

(a) Filter of transfer function $G[p] = \alpha_c/(\alpha_c + p)$
(b) Davis filter
(c) Filter of transfer function $G[p] = (1 - p^2)/(\alpha_c + p)^2$
(d) Synchronous motor of transfer function $G[p] = \alpha_c^2/(\alpha_c + p)^2$

If $\Delta N_1 = 1/1600$, $\alpha_c/2\pi = 1$c/s, $k_1 = \frac{1}{4}$, and $k_2 = \sqrt{2} - \frac{3}{4}$, one can find $\omega_1/2\pi$ as a function of m. This has been done and curve (b), Fig. 5 is a graph of the resulting function. It will be noticed that there is not a very great difference between the behaviour of the two types of filter. The Davis filter, designed so that $k_1 = 1$ and $k_2 = 2$, would behave as shown in curve (a), Fig. 5.

Mechanical Filter of Transfer Function $G[p] = 1 - p^2/(\alpha_c + p)^2$

T. C. Nuttall has pointed out to the author that filters of the general type

$$G_n[p] = 1 - p^n/(\alpha_c + p)^n \quad \ldots\ldots\ldots\ldots \quad (22)$$

are capable of giving rise to no more than a finite ultimate framing error after infinite frequency change, if the rate of change does not exceed

$$m = \omega_n/2\pi \left(1/T_1 + 2t/T_2^2 + 3t^2/T_3^3 + \ldots\ldots + \frac{(n-1)t^{n-2}}{T_{n-1}^{\,n-1}}\right)$$

where the T's are suitable constants. Such filters are therefore suitable for use in television film scanners or telerecording equipment because it is possible to restrict the maximum long-term framing error to any desired value by ensuring that the mains-hold shall not allow its output of field synchronizing pulses to change frequency at a rate exceeding a calculable tolerance. A simple practical example with $n=2$ will be examined. If, in Fig. 2 $R_2=0$, it is found that the Davis filter thus modified will have a transfer function of the desired form. Thus if $R_2 = 0$, $K_2 = K_1$ and making the further restriction that $K_1 = 2$, then equation (19) becomes

$$G_2[p] = 1 - p^2/(\alpha_c + p)^2 \quad \ldots\ldots\ldots\ldots \quad (24)$$

Putting $n = 2$ into equation (23), one finds that the rate of change of field frequency with respect to time must not exceed

$$m = \omega_0/2\pi T_1 \quad \ldots\ldots\ldots\ldots\ldots \quad (25)$$

In order to apply equation (13) since m is given by equation (25), $\omega(t)$ will be

$$\omega(t) = \omega_0 t/T_1 \quad \ldots\ldots\ldots\ldots\ldots \quad (26)$$

the constant of integration being omitted because $\omega/2\pi$ is regarded as the departure of field frequency from the nominal value of $\omega_0/2\pi$.

From equation (26)

$$\omega[p] = \omega_0/T_1 p \quad \ldots\ldots\ldots\ldots\ldots \quad (27)$$

and substituting equations (24) and (27) into equation (13)

$$\Delta N \simeq \omega_0/2\pi T_1(\alpha_c + p)^2 \quad \ldots\ldots\ldots \quad (28)$$

The 'd.c.' performance can be obtained by letting $p \to 0$; this allows one to obtain the ultimate framing error from equation (22) as follows:

$$\Delta N_{t\to\infty} = \Delta N_{p=0} \text{ or } \Delta N(\infty) = \Delta N[0]$$

whence

$$\Delta N(\infty) = \frac{\omega_0/\alpha_c}{2\pi\alpha_c T_1} \quad \ldots\ldots\ldots\ldots\ldots \quad (29)$$

or, substituting $\omega_0/2\pi T_1$ for m in equation (25)

$$\Delta N(\infty) = m/\alpha_c^2 \quad \ldots\ldots\ldots\ldots \quad (30)$$

The value $m = 0{\cdot}075$c/s/s, or 0·15 per cent of 50c/s in a second, allows an ultimate framing error of one line pitch in a 405-line television system if the mechanical filter just considered has a design frequency or nominal cut-off frequency of $\alpha_c/2\pi = 0{\cdot}88$c/s. The transient performance of this filter to the linear frequency change will now be examined. The right-hand side of equation (28) is the operational equivalent of

$$\Delta N = \frac{\omega_0/\alpha_c}{2\pi\alpha_c T_1}[1-(1+\alpha_c t)\exp(-\alpha_c t)] \quad \ldots\ldots \quad (31)$$

This function, like those studied in the previous sections, has no maximum, but unlike them, it tends to the finite limit given by equation (29). Curve (c), Fig. 5, shows how the limit is approached when expressed in terms of maximum permissible frequency deviation as a function of rate of change of frequency. Once again, $\Delta N_1 = 1/1600$ and $\alpha_c/2\pi = 1$c/s.

Mechanical Filter of Transfer Function $G[p] = \alpha_c^2/(\alpha_c+p)^2$

This transfer function can be shown to represent the ratio of the angular velocity $\omega_s = d\lambda/dt$ of the shaft of a synchronous motor to the angular frequency $\omega = d\theta/dt$ of the electric supply voltage applied to it, if the motor is suitably designed. Let the moment of inertia of the motor be I, the angular lag λ_0, the load (assumed to be viscous) ρ and the steady-state frequency of the supply voltage $\omega_0/2\pi$, then if

$$\left.\begin{aligned} \rho/I &= 4\omega_0 \tan\lambda_0 \\ \text{and } \alpha_c &= 2\omega_0/\tan\lambda_0 \end{aligned}\right\} \text{ motor design features}$$

it can be shown that

$$\lambda[p]/\theta[p] = \alpha_c^2/(\alpha_c + p)^2 \quad \ldots\ldots\ldots\ldots \quad (32)$$

which is also the ratio of the operational equivalents of shaft angular velocity to supply angular frequency.

Such a mechanical filter resembles those discussed previously in that, theoretically, an infinite ultimate framing error arises if the supply frequency changes linearly with time; however, synchronous motors are in use in some telerecording equipment and framing errors due to long-term drift in television field-pulse frequency can be corrected by manual adjustment. It is therefore desirable that the transient performance of such drive mechanisms be studied.

When $\omega[p] = 2\pi\, m/p$ and $G[p]$ takes the form given in equation (32) above, equation (13) becomes

$$\Delta N \simeq m\,[1 - \alpha_c^2/(\alpha_c + p)^2]/p^2 \quad \ldots\ldots\ldots \quad (33)$$

or

$$\Delta N = m/\alpha_c^2[-3 + 2\alpha_c t + (3+\alpha_c t)\exp(-\alpha_c t)] \quad \ldots\ldots \quad (34)$$

Curve 4, Fig. 5, shows the maximum permissible field-pulse frequency deviation as a function of rate of change for $\Delta N_1 = 1/1600$ and $\alpha_c/2\pi = 1$c/s.

THE FUNDAMENTAL COMPONENT OF THE MAINS-HOLD OUTPUT WAVEFORM HAS A FREQUENCY WHICH AS A FUNCTION OF TIME CONSISTS OF A SINGLE HALF-SINE WAVE

This assumption, expressed mathematically, is

$$\begin{aligned} &\omega(t) = \Delta\omega \sin\alpha t \text{ for } 0 \leqslant t \leqslant \pi/\alpha \\ &\omega(t) = 0 \text{ for } t \leqslant 0 \text{ and } t \geqslant \pi/\alpha \quad \ldots\ldots\ldots\ldots \quad (35) \end{aligned}$$

Using the delay operator, $e^{-\pi p/\alpha}$, one can express this time discontinuity by adding $\sin\alpha t$ to itself after a time equal to π/α; thus

$$\omega[p] = \Delta\omega\, \frac{\alpha p}{p^2+\alpha^2}(1 + e^{-\pi p/\alpha}) \quad \ldots\ldots\ldots \quad (36)$$

Mechanical Filter of Transfer Function $G[p] = \alpha_c/(\alpha_c + p)$

This transfer function and equation (36) can now be substituted into equation (13), thus obtaining

$$\Delta N \simeq \Delta\omega/2\pi \cdot \frac{\alpha}{p^2+\alpha^2}(1 + e^{-\pi p/\alpha}) \quad \left(1 - \frac{\alpha_c}{\alpha_c + p}\right). \quad (37)$$

The simplest way of solving the above operational equation is to let

$$\Delta N = \Delta\,\omega/2\pi\,[f(t) + f(t - \pi/\alpha)] \quad \ldots\ldots \quad (38)$$

with $f(t)$ valid for $t \geqslant 0$
and $f(t - \pi/\alpha)$ valid only when $t \geqslant \pi/\alpha$.

Using this method and Borel's product theorem,

$$\left.\begin{aligned} \Delta N &= 1/2\pi\,\frac{\Delta\omega/\alpha_c}{\sqrt{(1+\alpha^2/\alpha_c^2)}}\left[\sin(\alpha t - \arctan\alpha/\alpha_c) + \frac{\alpha/\alpha_c}{\sqrt{(1+\alpha^2/\alpha_c^2)}}\exp(-\alpha_c t)\right] \\ &\qquad \text{for } 0 \leqslant t \leqslant \pi/\alpha. \\ \Delta N &= 1/2\pi\,\frac{\Delta\omega/\alpha_c}{\sqrt{(1+\alpha^2/\alpha_c^2)}}\left[\frac{\alpha/\alpha_c}{\ } (1+\exp(\pi\alpha_c/\alpha))\exp(-\alpha_c t)\right] \\ &\qquad \text{for } t \geqslant \pi/\alpha. \end{aligned}\right\} \quad (39)$$

Unlike the examples dealt with previously, the expression in brackets, [ ], in equation (39) passes through a maximum, and if one finds, by drawing curves, (ΔN_1 and α_c being fixed), the values of this maximum for various values of α, one may evaluate $\Delta\omega$ in terms of α. Curve (a), Fig. 5, shows $\Delta\omega/2\pi$ as a function of $\alpha/2\pi$ for $\alpha_c = 2\pi$ rad/sec and $\Delta N_1 = 1/1600$.

The case of a Davis filter submitted to the same transient as the filter just treated becomes so unwieldy as to lose clarity and will not therefore be examined.

Mechanical Filter of Transfer Function $G[p] = 1 - p^2/(\alpha_c + p)^2$

For brevity, the calculations for this case are omitted. They are similar to those shown above. The maximum permissible deviation, $\Delta\omega/2\pi$, of the frequency of the field-pulse waveform, with $\Delta N_1 = 1/1600$, is shown as a function of the modulation frequency of the half-sine wave 'frequency pulse', $\omega(t) = \Delta\omega \sin \alpha t$, in Fig. 6, curve (d). As before, $\alpha_c/2\pi = 1$c/s.

Mechanical Filter of Transfer Function $\alpha_c^2/(\alpha_c + p)^2$

The transient response of the synchronous motor to a single half-sine wave frequency pulse may be obtained in a similar manner. Curve (e), Fig. 6, shows the maximum permissible frequency deviation, $\Delta\omega/2\pi$, of the field-pulse waveform as a function of the half-sine wave modulation frequency $\alpha/2\pi$, if $\Delta N_1 = 1/1600$ and $\alpha_c/2\pi = 1$c/s.

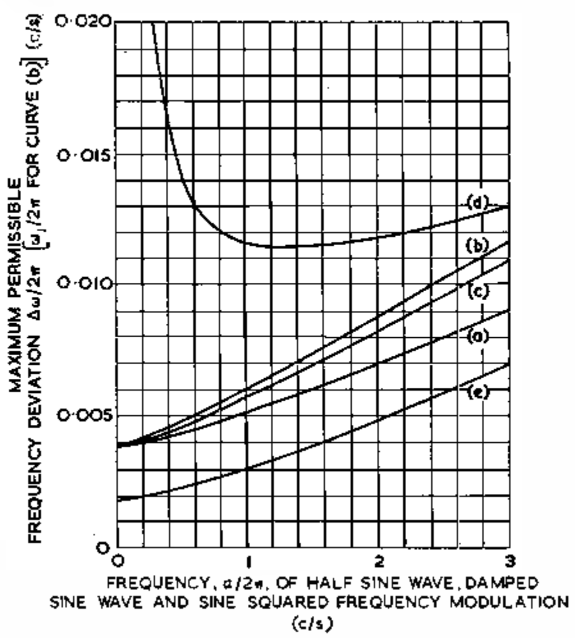


Fig. 6. Transient deviation of field frequency giving rise to a ½ line picture bounce: field frequency variation of pulse form

(a) Half sine wave frequency modulation
(b) Damped sine wave frequency modulation } *Filter*: $G[p] = \alpha_c/(\alpha_c + p)$
(c) Sine squared wave frequency modulation
(d) Half sine wave frequency modulation. Filter: $G[p] = (1 - p^2)/(\alpha_c + p)^2$
(e) Half sine wave frequency modulation. Filter: $G[p] = \alpha_c^2/(\alpha_c - p)^2$

THE FUNDAMENTAL COMPONENT OF THE MAINS-HOLD OUTPUT WAVEFORM HAS A FREQUENCY WHICH AS A FUNCTION OF TIME CONSISTS OF A SINE-SQUARED PULSE

For this one has

$$\omega(t) = \Delta\omega \sin^2 \alpha t \text{ for } 0 \leqslant t \leqslant \pi/\alpha$$
$$\omega(t) = 0 \text{ for } t \leqslant 0 \text{ and } t > \pi/\alpha \quad \ldots\ldots\ldots (40)$$

Using the delay operator, $e^{-\pi p/\alpha}$, one can write immediately

$$\omega[p] = \Delta\omega \frac{2}{4 + p^2/\alpha^2}(1 - e^{-\pi p/\alpha}) \quad \ldots\ldots\ldots (41)$$

Mechanical Filter of Transfer Function $G[p] = \alpha_c/(\alpha_c + p)$

Substituting this transfer function and equation (41) into equation (13) and using the same methods as in previous sections, curve (c), Fig. 6 is obtained, for which let $\alpha_c = 2\pi$ rad/sec and $\Delta N_1 = 1/1600$.

THE FUNDAMENTAL COMPONENT OF THE MAINS-HOLD OUTPUT WAVEFORM HAS A FREQUENCY WHICH AS A FUNCTION OF TIME CONSISTS OF A DAMPED SINE WAVE

For this case

$$\omega(t) = \Delta\omega\, e^{-at} \sin \alpha t \text{ for } t \geqslant 0 \quad \ldots\ldots\ldots (42)$$

and

$$\omega[p] = \Delta\omega \frac{\alpha p}{(p + a)^2 + \alpha^2} \quad \ldots\ldots\ldots (43)$$

Mechanical Filter of Transfer Function $G[p] = \alpha_c/(\alpha_c + p)$

Substituting this transfer function and equation (43) into equation (13),

$$\Delta N \simeq \Delta\omega/2\pi \cdot \frac{\alpha}{(p + a)^2 + \alpha^2}\left(1 - \frac{\alpha_c}{\alpha_c + p}\right) \ldots (44)$$

and finally,

$$\Delta N = 1/2\pi \cdot \frac{\Delta\omega/\alpha_c}{\sqrt{[(1 - a/\alpha_c)^2 + \alpha^2/\alpha_c^2]}}\left[\exp(-\alpha_c t) \sin\phi + e^{-at} \sin(\alpha t - \phi)\right] \quad (45)$$

where $\phi = \arctan \dfrac{\alpha/\alpha_c}{1 - a/\alpha_c}$

For simplicity and because good mains-hold design would probably ensure it, let $a = \alpha$ in the following example wherein $\alpha_c = 2\pi$ rad/sec and $\Delta N_1 = 1/1600$. Table 1 shows the time, t_1, of occurrence of the first maximum of expression (45) as well as the permissible value of $\Delta\omega/2\pi$ and the magnitude of the maximum permissible frequency deviation

$$\omega_1/2\pi = \Delta\omega/(2\pi\sqrt{2}) \cdot e^{-\pi/4} \quad \ldots\ldots\ldots (46)$$

for various values of $a = \alpha$.

TABLE 1.

$a = \alpha$ (rad/sec)	t_1 (sec)	$\Delta\omega/2\pi$ (c/s)	$\omega_1/2\pi$ (c/s)
$\ll a_c$	$1/\alpha \arctan \alpha/a$	0·0121	0·00390
1	1	0·0125	0·00403
1·5	0·7	0·0129	0·00416
2	0·55	0·0135	0·00435
2π	0·255	0·0188	0·00606
10	0·18	0·0237	0·00765
20	0·10	0·0364	0·0117

It should be noted that curve (b), Fig. 6, shows $\omega_1/2\pi$ as a function of $\alpha/2\pi$ and not, in this case, $\Delta\omega/2\pi$ as the latter is not the maximum permissible frequency deviation (which is what is shown as the ordinate of Fig. 6) but merely the maximum permissible coefficient. Note also that although the above results and curve (b), Fig. 6, only apply to a filter of transfer function $\alpha_c/(\alpha_c + p)$ the Davis filter can be made to have just this transfer function.

The mains-hold transient waveform dealt with and the results obtained in the last section have considerable practical significance since a typical modern mains-hold circuit is found to respond to a unit-step change in mains phase in the way that has been discussed.

A Practical Example

Suppose one has a mains-hold consisting of a phase discriminator so designed as to produce an output voltage proportional to the phase difference between the mains supply and the fundamental component of field-synchronizing-pulse waveform. Let this output voltage control the frequency of the field-synchronizing-pulse waveform in a linear manner by means of a suitable reactance-valve circuit, and assume that the long time-constant capacitance-resistance-capacitance circuit shown in Fig. 7 is interposed between the phase-discriminator output and the reactance valve*. Using the same notation as before, Fig. 8 shows a schematic diagram of the mains-hold components.

The voltage appearing at the output of the phase discriminator will be, if β_1 is a suitable constant,

$$V_1 = \beta_1 (\theta_m - \theta) \quad \ldots\ldots\ldots\ldots\ldots \quad (47)$$

while the voltage, V_2, at the input to the reactance valve will be given by

$$V_2[p] = 8g_m/C \;.\; 1/p \;.\; \frac{1 + \tau p}{9 + \tau p} \; V_1[p] \quad \ldots \quad (48)$$

where $\tau = RC$. If the reactance valve-cum-oscillator combination is linear, one may write

$$\omega = \beta_2 V_2 + \omega_0 \quad \ldots\ldots\ldots\ldots\ldots \quad (49)$$

if β_2 is a suitable quantity (for example, of dimensions: radians per second per volt).

Substituting angular frequencies for phase angles in equation (47), one obtains from equations (47), (48), (49)

$$\omega[p] = 27/\tau^3 (1 + \tau p)/(3/\tau + p)^3 \, \omega_m[p] \quad \ldots \quad (50)$$

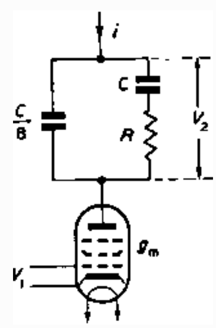
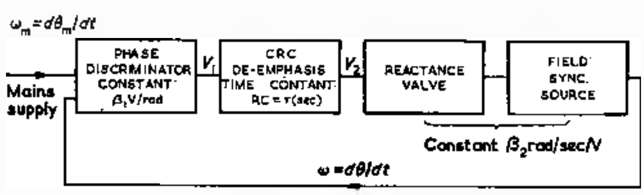


Fig. 7. Smoothing circuit with infinite gain for d.c.

Fig. 8. The mains-hold components



in which

$$8g_m\beta_1\beta_2/C = 27/\tau^2$$

Now assuming that the disturbance in the mains supply consists of a unit-step change of phase, θ_0. Then

$$\omega_m[p] = \theta_0 p \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (51)$$

Substituting equation (51) into (50) and re-arranging

$$\omega[p] = 27\theta_0 p/\tau^3 (1 + \tau p)/(3/\tau + p)^3 \quad \ldots \quad (52)$$

whence

$$\omega(t) = 27\theta_0 t(1 - t/\tau)\, e^{-3t/\tau} /\tau^2 \quad \ldots\ldots \quad (53)$$

This function is shown as curve (a), Fig. 9. It is not surprising that it is similar to the damped sine wave of equation (42), if it is arranged that the damping has the critical value $a = \alpha$. Curve (b), Fig. 9, shows the variations of frequency in accordance with equation (42) when $a = \alpha = \pi/\tau$ and $\Delta\omega = 7{\cdot}48\,\theta_0/\tau$. In a certain mains-hold τ has been given the value $\tau = 100$sec. Thus referring to curve (b), Fig. 6, one finds that for $\alpha/2\pi = 0{\cdot}005$ $[= \pi/(2\pi\tau)]$, $\omega_1/2\pi = 0{\cdot}0039$c/s and referring to Table 1, Column 3, $\Delta\omega/2\pi = 0{\cdot}0121$c/s. If such a mains-hold were employed in the waveform generator of a television transmitting system transmitting a film by a continuous-motion scanner containing a mechanical filter of transfer function $\alpha_c/(\alpha_c + p)$, the resulting image as seen on television receiver screens would bounce by a quarter of a line pitch in the frame direction if the mains waveform underwent a unit-step change of phase of $\theta_0 = \tau\Delta\omega/7{\cdot}48$ or $\theta_0 = 1$rad. This is a most interesting result because the largest possible effective mains phase shift must be π radians. Thus it is seen that a mains-hold of the type described and having the time-constant taken as an example would be most unlikely to give rise to picture bounce exceeding a quarter of a line pitch unless film scanning and recording machines had mechanical filters with cut-off frequencies much less than $\alpha_c/2\pi = 1$c/s.

Conclusions

The response of a number of different types of mechanical filter to several forms of excitation representative or typical of the type of waveform to be expected from a mains-hold circuit when subjected to a transient of phase in the electricity supply mains have been examined, and it should now be possible to formulate a method of specifying the tolerances on the transient performance of mains-hold circuits to be used at television signal sources.

First, the specification of a maximum rate of change of frequency, (Fig. 5) alone, is insufficient to be certain of avoiding undesirable picture bounce. Since it is unlikely that a mains-hold will be designed whose response to a mains-phase transient will have a constant rate of change

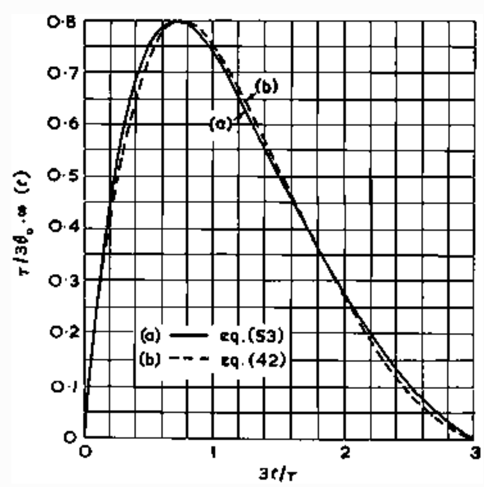


Fig. 9. Variation of frequency of the waveform of the output from a mains-hold with wide-range phase detector when a unit-step change of phase occurs in the electric supply mains

of frequency, it does not seem to be a sufficient method of mains-hold specification.

Secondly, it has been shown that at least for one type of mains-hold circuit it is possible by suitable choice of time-constant to ensure that no normally encountered mains phase transient is likely to cause an intolerable picture bounce.

Finally, it will now be evident that it is not strictly possible to specify the performance required from a mains-hold unless the characteristics of the device having the largest time-constant in the television service are known as well as those of the mains-hold itself; that is, one must know both $G[p] = G[j\alpha]$ and $\omega[p] \simeq \omega(t)$.

One might, therefore, specify to the designer of a waveform generator a maximum permissible picture bounce and inform him of the transfer function, $G[p]$, of say, the film-recording film-traction mechanism or even the film-traction mechanism of a typical continuous motion film scanner—in which case the designer himself will have to choose the type of mains-hold and then study its behaviour as has been done in the example—or one can take a typical filter transfer function and pre-suppose that the designer will design a mains-hold whose performance will not differ

* *The desirability of using this particular circuit because of its infinite response to a sinusoidal excitation of zero frequency was pointed out to the author by E.L.C. White, who kindly made available some unpublished work by W.S. Percival and himself.*

too widely from the functions $\omega(t)$ already discussed. In this case it would be possible to issue a curve such as any one of those in Fig. 6 and the specification would, quite simply, take the form of a prohibition that any combination of α and $\Delta\omega$ should define a point above the curve. This would protect the viewer from bounce but not from long-term framing errors; however, the latter can be corrected manually at the television studio, although protection from long-term framing errors is considered to be highly desirable.

In spite of the apparent simplicity of the latter method of specification, it is considered that a more precise method would be to take the simplified synchronous motor transfer function on account of its general utility and specify the transient performance of the mains-hold by reference to equation (13). With regard to the question of rate of change of field synchronizing frequency, this applies more directly to the mechanical filter performance than to that of the mains-hold since there seems to be no *a priori* reason for supposing that the mains-hold should change frequency faster than the electricity supply mains, (0·001c/s/s as quoted at the beginning of this article).

If one substitutes the transfer function of the synchronous motor from equation (32) into equation (13) and express ΔN as a convolution integral (Borel's theorem), one may limit the transient picture bounce to $\frac{1}{4}$ line pitch of a 405-line 50-field television system by writing

$$2\pi/1600 \geqslant |\exp(-\alpha_c t)\Big\{(1+\alpha_c t)\int_0^t \omega(x)\exp(\alpha x)\,dx - \alpha_c \int_0^t x\omega(x)\exp(\alpha_c x)\,dx\Big\}\max \quad \ldots . (54)$$

in which one might let $\alpha_c = 2\pi$ rad/sec and where $\omega(t)$ is the transient response of the mains-hold in question to a specified unit-step change θ_o of mains phase. A value $\pi/2 < \theta_o < \pi$ would not seem unreasonable.

Acknowledgments

The author is indebted to Dr. E. L. C. White and Mr. T. C. Nuttall for their very helpful comments and contributions to the theoretical work which gave rise to this article. Thanks are due to the Chief Engineer of the BBC for his kind permission to publish this artcile.

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A One-Valve D.C. Amplifier with High-Impedance Input

By P. Belton*, B.Sc., A.R.C.S.

This article describes a very simple probe for use in physiological research primarily for measuring steady intra-cellular potentials. It is intended for direct connexion to a commercially available oscilloscope. Drift in the first half hour of use seldom exceeds 1mV. A simple compensating circuit is also described which has the effect of improving the input capacitance so that action potentials may be satisfactorily rendered.

WHERE high-impedance electrodes are used in electrophysiological research, as in recording membrane-potentials with intra-cellular micro-electrodes some form of impedance transformer is needed before the signals can be applied to orthodox valve amplifiers, and many circuits have been developed with this in view.

With a suitable choice of valve, the impedance transformation, together with a stage gain of about 15, should be attainable using only a single valve in a simple circuit such as that used in the Cossor 3327 d.c. amplifier (1938), with certain modifications.

Specification

The general requirements† which have to be met by apparatus used in this field may be set out as follows:

Input Impedance

Leakage resistance at the input terminal should be large enough to be disregarded when compared to the input grid current. The grid current should not be greater than 10^{-11}A, (this is small enough to avoid stimulation of normal nerve and muscle cells).

Rise-time

The input time-constant should not be greater than 30μsec. This is the time-constant recommended by Nastuk[2] for the recording of muscle end-plate potentials with negligible distortion on the basis of comparisons with the results of graphical analysis.

Sensitivity

The overall sensitivity at the tube face of the oscilloscope should be about 10mV/cm.

Stability

The apparatus should be able to measure steady potentials in the order of 50mV for periods of half an hour without drifting more than 1 per cent.

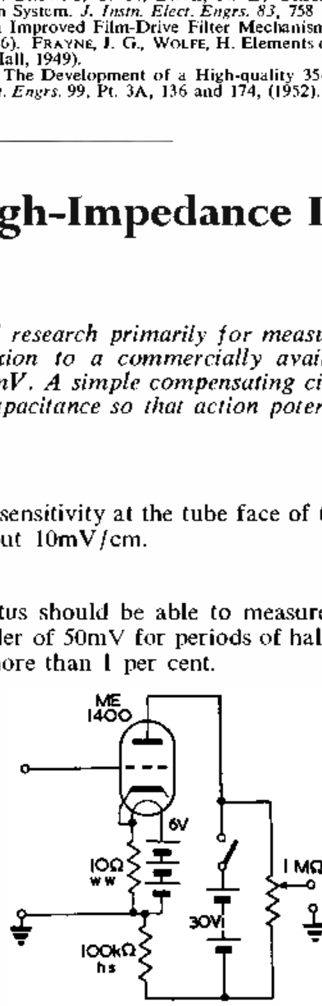


Fig. 1. Circuit of D.C. amplifier

Input Level

The input terminal should be near earth potential to avoid stimulating the preparation.

The one valve amplifier described conforms to this specification, except in regard to rise-time, when used with the Cossor 1049 oscilloscope. A simple correcting circuit is given, by which the input time-constant may be reduced to about 50μsec.

* *Zoology Department, Glasgow University.*

† *A more detailed specification for the wide range of instruments used for electrobiological measurement has been drawn up by Grundfest[1].*

Circuit

The circuit of the amplifier is given in Fig. 1. It comprises a triode amplifier whose anode load resistor and h.t. battery have been interchanged in order to obtain an output near earth level. Adjustment of output voltage-level is made after the apparatus has warmed up by adjusting the 1MΩ potentiometer, (since it was found impossible to tap the required voltage satisfactorily from the layer type battery used). Any later movement of the trace is made with the Y_1 shift control of the oscilloscope in order to prevent a change in the series resistance of the potentiometer affecting the gain of the valve.

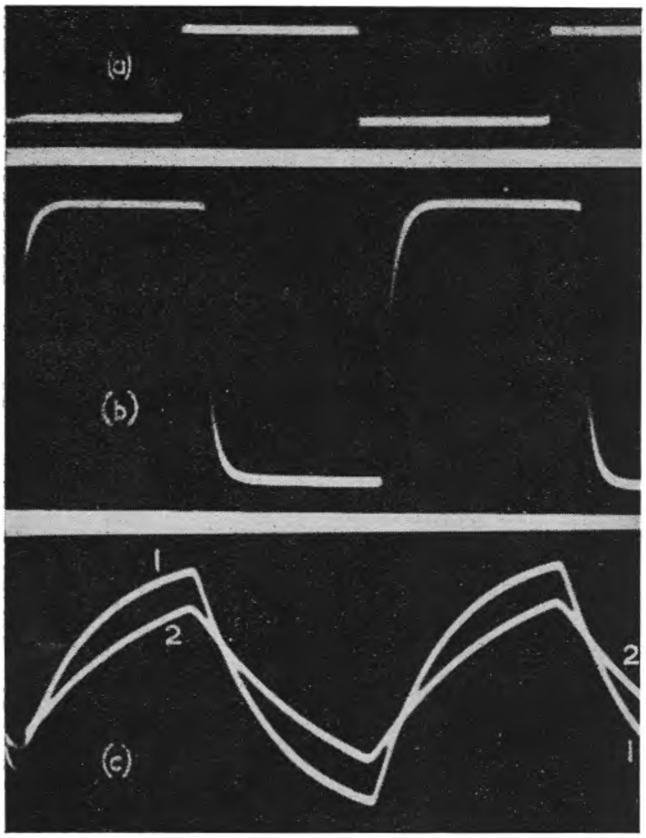


Fig. 2. Photographs of traces obtained with the Cossor 1049 oscilloscope
(a) 1kc/s square-wave applied direct to oscilloscope
(b) Same square-wave applied to input grid of d.c. amplifier
(c) Same square-wave applied to d.c. amplifier through dummy electrodes, consisting of a series resistance together with a shunt capacitance between grid and earth:
1 10MΩ and 2pF; 2 20MΩ and 2pF

An ME1400 semi-electrometer valve is used, triode connected. Stage gain, working with a 30V hearing-aid battery and an anode load resistor of 100kΩ, is in the region of 5, producing adequate deflexion on the Cossor 1049 oscilloscope. Grid current is kept to a low value (10^{-11}A) by applying a fixed bias between grid and cathode. The bias voltage is not presented across the preparation.

Response to Transients

The h.t. battery is separated from the chassis by a space of ⅛in since it changes potential with respect to earth by the magnitude of the amplified signal. However, with an anode load of 100kΩ the time-constant of this part of the circuit is only some 10μsec; this may be seen from Fig. 2(b) which shows the response to the 1kc/s square-wave applied to the grid of the valve. The time-constant present in the grid circuit during normal operating conditions is much larger owing to the high series resistance of the micro-electrode. The capacitance affecting the input time-constant is made up of three components: (a) that due to the valve, which can be expressed as:

$$C_i = C_{ag} + C_{gk}(1 + A)$$

where A is the amplification factor; (b) the grid circuit stray capacitances; and (c) the capacitance due to the thin glass walls of the micro-electrode where it is immersed in an earthy conducting liquid. With a 10MΩ electrode in circuit, the total time-constant for the apparatus is in the order of 250μsec. This may be seen from Fig. 2(c) which shows the response of the apparatus to the 1kc/s square-wave applied through a dummy electrode consisting of a series resistor and shunt capacitor connected to the input grid. The large time-constant is no disadvantage when all that is required is a measurement of the resting membrane-potential with some indication of the presence or absence of action-potentials. A time-constant of this magnitude would be a serious disadvantage if one wished to measure the size or rate of rise of action-potentials. Considerable improvement can, however, be achieved with a simple correcting circuit.

Correcting Circuit

It will be seen from Fig. 3 that the resistance of the

Fig. 3. Schematic diagram of amplifier and micro-electrode

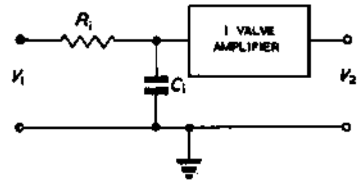
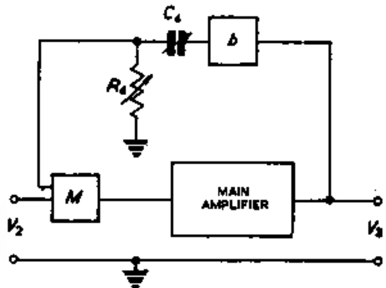


Fig. 4. Schematic diagram of correcting circuit
M mixing stage, b: attenuator in feedback loop



micro-electrode R_i, together with C_i the total shunt capacitance referred to above, form an integrating circuit of time-constant τ_1, which is presented to the input signals. The input voltage V_1 and the output voltage V_2 of the amplifier can be related by the expression:

$$V_2 = AV_1 \frac{1}{1 + j\omega\tau_1} \qquad (1)$$

where A is the gain of the amplifier.

The distortion represented by this expression can be corrected in several ways by the application of positive feedback. Haapanen and Ottoson[3] increase the gain of their cathode-follower to unity and neutralize the capacitance existing between grid and cathode. Solms, Nastuk and Alexander[4] use cascaded cathode-followers to reduce the input capacitance, and reduce it still further by applying the output of a 'negative capacitance amplifier' to the input grid. Woodbury[5] applies correction at a later stage by differentiating the distorted signal, the differentiated signal is then amplified and added to the original. The simple method of reducing distortion used in the present instance is based on the principle used by Woodbury. In this case a proportion of the signal obtained from a differentiating circuit connected to the output of the main amplifier is fed back to the input of the amplifier. A schematic diagram is given in Fig. 4. The author is indebted to Mr. A. M. Andrew for an analysis of this type of circuit, which does not assume mathematically perfect differentiation, as Woodbury's seems to have done.

With reference to Fig. 4, the main amplifier has a gain A', the positive feedback loop has a gain b, and the differentiating circuit has a time-constant of τ_2. The voltage V_3 at the tube face of the oscilloscope is related to the voltage V_2 at the input terminal of the oscilloscope by the expression:

$$V_3 = A'\left(V_2 + b\,\frac{j\omega\tau_2}{1 + j\omega\tau_2}\,V_3\right)$$

$$\therefore\ V_3\left(1 - \frac{A'bj\omega\tau_2}{1 + j\omega\tau_2}\right) = A'\,V_2$$

$$\therefore\ V_3\,\frac{1 + (1 - A'b)\,j\omega\tau_2}{1 + j\omega\tau_2} = A'\,V_2$$

Substituting for V_2 from equation (1),

$$V_3 = A'A\,\frac{1 + j\omega\tau_2}{(1 + j\omega\tau_1)\,[1 + (1 - A'b)\,j\omega\tau_2]}\,V_1$$

Now, if $A'b = 1$ (critical feedback), and if τ_2 is made equal to τ_1 this equation reduces to $V_3 = A'A\ V_1$, so that there is no distortion. In practice a perfect correction cannot be made, on account of the phase shifts in the amplifiers, and the fact that, with critical feedback, the circuit oscillates since the overall gain of the system in this condition is infinite.

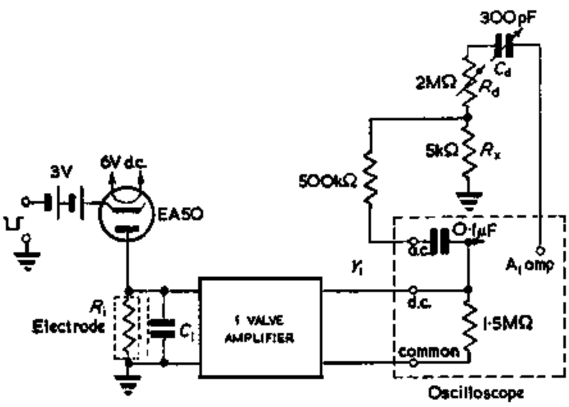


Fig. 5. Diagram of complete apparatus

The circuit in use is shown in Fig. 5. The differentiating circuit consists of C_d with R_d and R_x. Attenuation (corresponding to the gain b of the feedback loop) is carried out in two stages by potential dividers: firstly by R_d with R_x, and secondly by the isolating resistor with the effective input resistance of the oscilloscope. It can be shown that the output impedance of the oscilloscope amplifier can be neglected, as there is no change in performance of the system when a cathode follower is inserted between the output of the oscilloscope amplifier and the differentiating network.

Setting-Up

The resistance of a micro-electrode used for measuring resting and action-potentials may lie between 5 and 20MΩ. The time-constant of the correcting circuit must therefore be adjusted to suit the electrode in use. The method used by Haapanen and Ottoson has been found best in practice. The principle of their method is to charge the input capacitance and allow it to discharge through the resistance of the electrode, which is connected up with its tip dipped into earthy saline for 1 to 2mm in order to simulate recording conditions. Negative-going pulses from a low impedance generator are applied to the input grid through a battery in series with a thermionic diode. The capacitance is charged rapidly through a low impedance during that part of the pulse which exceeds the battery voltage. When the applied voltage starts to move positively, the diode presents its very large reverse resistance and the input capacitance can only discharge through the electrode to give a lengthened exponential tail. The input time-constant can be readily observed from this waveform, and R_d,C_d are adjusted until it is as square as possible. The rise-time of the apparatus can be reduced in this way to between 30 and 50μsec depending on the resistance of the electrode, this can be seen from Fig. 6 which shows the corrections obtainable using electrodes of 5 and 20MΩ.

Construction

The amplifier is built around the valve base of the ME1400, and enclosed in a steel screening box. This is clamped over the micro-electrode and a short length of chlorided silver wire, connected to the top-cap of the inverted valve, is dipped into the 3 molar potassium chloride which fills the electrode.

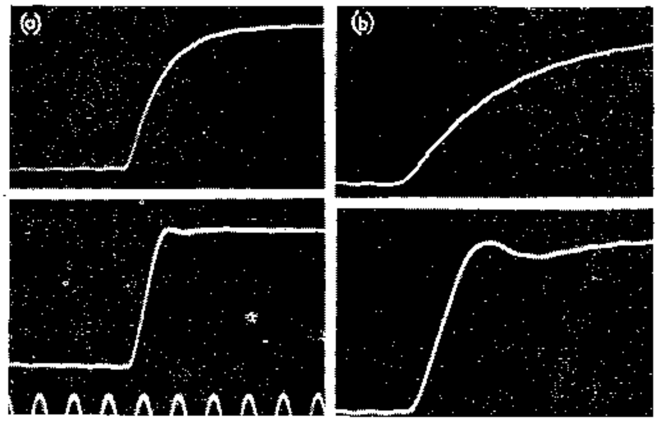


Fig. 6. Effect of correcting circuit

(a) Square pulse applied to micro-electrode of 5MΩ. Above, uncorrected; below, corrected; timing wave 20kc/s.

(b) Same pulse applied to micro-electrode of 20MΩ. Above, uncorrected; below, slightly over-corrected.

The correcting components are mounted on a paxolin sheet at the left of the oscilloscope. There has been no trouble with hum pick-up in this part of the circuit and no screening has been used since stray capacitance to earth reduces the efficiency of the correcting circuit.

Conclusion

While no great claims are made for the fidelity of this apparatus, it is hoped that its compact size, simplicity and stability, and perhaps most important, its low cost will enable the use of capillary micro-electrodes to become more general in physiological research.

Acknowledgments

The author is indebted to Dr. T. D. M. Roberts for his kind help with the script, and to A. C. Cossor Limited, for the supply of the manual for their d.c. amplifier type 3327 and for permission to utilize material from it. The work was carried out while the author was receiving a maintenance allowance from the Department of Scientific and Industrial Research.

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Short News Items

Ferranti Ltd have announced that two more electronic digital computers have gone into service in the British aircraft industry. A Pegasus Computer has been installed and is now working at the Brough Factory of Blackburn and General Aircraft Ltd. It is employed on design work and performance study for the Blackburn NA.39 Strike Aircraft. Considerable work for the NA.39 has already been completed by the Pegasus Computer at the London Computer Centre and on another Pegasus at Leeds University. The work at Brough will include all aspects of aeronautics such as aerodynamics, structures, aerolastics, test flight and wind tunnel data reduction, basic geometric design and data for tangent planing of wind tunnel models. The Pegasus will be employed not only on calculations for the Aircraft Design of Blackburn's but also for the Engine Division. The second computer, a Ferranti Mark 1 at Armstrong Siddeley Motors, Ansty, is working on engine performance calculations, prediction of compressor and turbine characteristics, whirling speeds and disk stressing. This is the fourth computer to be installed in the Hawker Siddeley Group and brings the total number of computers in service in the British aircraft industry to fifteen.

The Development and Workshop Production sections of Cable and Wireless Ltd Engineer-in-Chief's Department at Radio House are to move to new premises in Kirby Street, London, E.C.1. About 100 personnel are involved in the move. The new building has been named Smale House, after Mr. J. A. Smale, Engineer-in-Chief of Cable and Wireless Ltd until he retired in March 1957, who was closely associated with the research and development work of the Company.

Marconi's Wireless Telegraph Co Ltd have received an order from Jordan for a high-power transmitter for a new radio station to be built in Amman. It is planned to bring this station on the air early in 1959. Marconi's are to supply a 100kW medium wave transmitter Type BD206. This air-cooled transmitter is one of the range of high power broadcast equipment designed by the Company to provide high fidelity transmission. Two of the main advantages of this transmitter are its low running costs and simplicity of operation.

The BBC has announced that Mr. F. C. Brooker has been appointed Engineer-in-Charge, London (Sound) in succession to Mr. C. E. Bottle, M.B.E., who has retired after thirty-four years' service.

Standard Telephones and Cables Ltd have now produced a numerical display indicator tube which has wide applications wherever a direct digital readout is required. This is a cold cathode device having ten cathodes each being shaped in the form of a numeral (0-9) and brought out to a separate pin on a duodecal base. The application of approximately 200V between a selected cathode and the common anode causes a glow discharge in the form of the appropriate numeral. The cathodes are stacked parallel to the valve base and the illuminated cathode is viewed through the top of the glass envelope. The seated height of the valve is $1\frac{7}{8}$in and bulb diameter $1\frac{1}{4}$in. The size of the numerals is approximately $\frac{7}{8}$in. They can be read clearly at a distance of 40-50ft.

Marconi Instruments Ltd have reported sales amounting to nearly £40 000 for the Marconi Television Transmitter Sideband Analyser Type OA 1241. Since details of this unique test equipment were released a few months ago, twenty-seven of the instruments have been ordered. The BBC and ITA networks are being equipped with the OA 1241; the instrument has already been delivered to a number of BBC stations, also to Denmark and Poland. Sweden, Switzerland and Brazil have also ordered the equipment, and enquiries have been received from many other countries.

Racal Engineering Ltd, of Bracknell, Berkshire, are now acting as exclusive agents in the United Kingdom for the sale of Transitron equipment manufactured by Van Norman Industries Inc of Manchester, New Hampshire, U.S.A. Included in the Transitron range are v.h.f. and u.h.f. signal generators and sweep generators of most advanced design. Technical details and prices of these equipments can be obtained from Racal Engineering Ltd.

Aero Research Ltd of Duxford, Cambridge, have announced that the name of the Company has been changed to CIBA (A.R.L.) Ltd. The trade names of 'Aerolite' 'Araldite', 'Redux', and other proprietary products are unaffected.

E.M.I. Electronics Ltd have received orders from the British Broadcasting Corporation valued at over £16 000, for the Company's type WM.2 television oscilloscopes, line selectors and ancillary units. This version of the WM.2 oscilloscope has been specially designed for monitoring television waveforms, and will be used by the BBC at transmitting stations, switching centres, studios and outside broadcast units. Special consideration has been given in the design to ensure that the instrument meets all essential measurement requirements associated with sine squared pulse and bar testing techniques which are now being widely accepted as a standard method for rapidly checking the characteristics of video transmission lines and networks.

The General Electric Co Ltd have announced the appointment of Mr. J. Bell as Manager of the Research Laboratories at Wembley. Mr. Bell will take charge of the administration of the group of establishments, the responsibility for the scientific policy and programmes remaining with the present Director of Research, Mr. O. W. Humphreys, C.B.E., who is also Director of the Applied Electronics Laboratories of the Company at Stanmore. The work of the Wembley Laboratories has been organized since 1953 in three divisions: Physics and Applied Physics, Chemistry and Engineering, and Telecommunications. Mr. Bell has been Manager of the Telecommuncations Division since its formation and retains the leadership of this work in addition to his new responsibility.

Metropolitan-Vickers Electrical Co Ltd have announced the appointment of Dr. L. W. Brown as Assistant Chief Electrical Engineer (Light Current), in addition to his present duties as Chief Engineer, Electronics Department.

The Guild of Air Traffic Control Officers will be holding a Convention at Southend on 23 and 24 October. The programme will be on similar lines to the last convention held in 1956. There will be talks by recognized authorities on the various aspects of air traffic control and general discussions. Further information may be obtained from Mr. G. L. Chambers, Clerk of the Guild of Air Traffic Control Officers, 118 Mount Street, Berkeley Square, London, W.1.

The Order of Knight Bachelor has been conferred on Professor Willis Jackson in the Birthday Honours list; he is director of Research and Education, Metropolitan-Vickers Electrical Co Ltd.

Erratum. Equation (13) of the article, "A Method for Sharpening the Output Waveform for Junction Transistor Multivibrator Circuits", by A. E. Jackets, in the June issue, should read:—

$$t_1 = C_1 R_1 \ln \left[\frac{V_{cc} + V_{bb} + I_{co}R_1}{V_{bb} + I_{co}R_1} \right]$$

LETTERS TO THE EDITOR

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

A Cathode-Ray Tube Function Generator

DEAR SIR,—A well-known type of function generator employs a cathode-ray tube and photo-electric cell with a mask interposed between them. The edge of the mask is cut to the shape of the function to be generated. If use is made of the secondary emission current from the screen as discussed in the recent articles[1,2] in the *Journal of Scientific Instruments*, the photo-electric cell in the above arrangement may be dispensed with. In this system, the secondary emission current from the screen is influenced by the potential of the mask which is in the form of a metal electrode in contact with the outside of the screen (Fig. 1).

separated by a small gap along the curve.

For a field voltage of some 20 or 30V the output across the collector resistor will be of the order of 10mV, so that the gain of the amplifier and detector need only be of the order of a few thousand.

If electrostatic deflexion is used, then the deflecting plates will tend to collect some of the primary beam current. This will result in a reduction in the secondary emission current, and with this type of deflexion the sensitivity of the device will vary over screen surface, although satisfactory working may be obtained over an appreciate part of the screen. If magnetic deflexion is employed the sensitivity will be more or less constant over the entire screen.

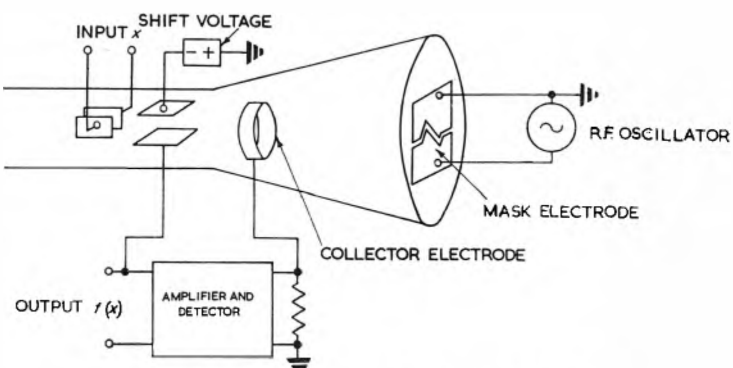


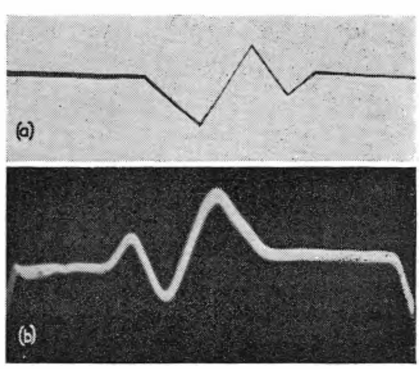
Fig. 1. General arrangement of the function generator

The electrode is energized by an oscillator with a frequency of about 500kc/s. A signal at this frequency is developed across a resistor in series with the electrode which collects the modulated secondary emission current when the electron beam falls on a point under the mask. The signal is amplified, demodulated and then, after further amplification applied to the vertical deflexion plates of the oscillograph. The polarity of this signal is such that it tends to deflect the beam away from the mask. As the spot moves from under the mask this signal is reduced to zero. In order to keep the signal at the edge of the mask a steady signal is applied at all times to the vertical plates in such a direction as to accomplish this, i.e., downwards if the mask covers the lower half of the tube face. If now the independent variable is caused to deflect the spot in the horizontal direction, the spot will move along the edge of the mask and the voltage on the vertical plates will be proportional to the function to be generated.

A satisfactory electrode may be constructed from a sheet of aluminium foil which has the required curve cut across it. The resulting two pieces of foil are then fixed to the face of the tube and

A limitation is imposed on the accuracy of the device by the fact that the electrodes defining the curve are separated from the screen surface at which the secondary emission takes place by an appreciable thickness of glass. The fringing[2] of the field as it spreads through the glass results in the path traced by the spot differing slightly from that defined by the edge of the mask. The effect is only serious when there are sharp changes in direction in the curve, when the effect of the fringing of the field will

Fig. 2. (a) Mask electrode
(b) Output voltage



be to cause the generated curve to be rounded at these corners. The effect can be seen in the trace reproduced in Fig. 2. This curve was obtained using a Cossor double beam tube, type 09, which is a very convenient tube for this purpose since it has an easily accessible collector and anode, A_4. A disadvantage of the tube, however, is that since the splitter plate used to produce the two beams interrupts an appreciable amount of current over certain areas of the screen this greatly reducing the sensitivity in this area.

The accuracy of the device is not great primarily because of the need to use only the centre area of the screen which limits the size of the mask electrode and again the fringing effect distorts the field pattern on the inside of the screen. In spite of this, it compares favourably with the similar system of photocell and mask since it has approximately the same size of mask and the photocell and amplifiers are replaced by a simple oscillator, amplifier and detector.

Yours faithfully,

J. WILLIS,

The Manchester College of Science and Technology.

REFERENCES

1. BURTT, R. B., WILLIS, J. A High Speed Electronic Analogue Field Mapper. *J. Sci. Instrum.* *34*, 177, (1957).
2. A Cathode-ray Tube Analogue-to-serial Digital Convertor. (To be published, *J. Sci. Instrum.*).

An Internal Defibrillator with Current Measuring Facilities

DEAR SIR,—While admiring the ingenious circuit of the defibrillator described by Dr. Perry and Mr. Trotman in the January issue, I would like to register a protest at the unnecessary and indeed dangerous complexity of the resulting instrument. If ventricular fibrillation should occur in the operating theatre, the utmost promptitude is necessary in treatment, every minute's delay decreasing the probability of success. Time to warm up the apparatus or to study its operation is not available. Further, the resistance of the human heart varies little from 50Ω, so that if 110V r.m.s. is applied, an adequate shock is given, and no improvement has been found by varying the current. Certainly there is no advantage in knowing the current after the shock has been administered!

A design devised in this department, which is in use in every large hospital in Melbourne, is shown in Fig. 1. To operate it, it is necessary only to connect it to the mains and depress the foot switch as required.

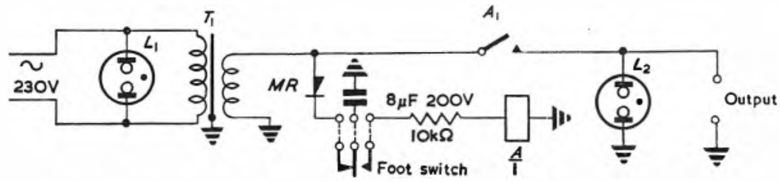


Fig. 1. A simple cardiac defibrillator

L_1 is a 230V neon lamp; L_2 is a similar lamp, but capable of striking on 100V r.m.s. T_1 is a standard doubly wound 230V to 110V 200VA step down transformer. $A/1$ is a GPO type 3 000 relay of 5kΩ, with a heavy-duty contact. The foot switch and mains flexibles are permanently attached and stored inside the lid of the instrument, while patient electrodes are maintained sterile in a polythene bag in the same compartment. The constants shown give a duration of 0·5sec. which has been found optimum.

Yours faithfully,

D. J. DEWHURST,

Department of Physiology,
University of Melbourne.

The authors' reply :

DEAR SIR,—Mr. Dewhurst need not worry, our circuit is simple at 'heart.' If it is so required, a pulse can be delivered 'blind' immediately our defibrillator is plugged in and switched on. Such rush action is, however, not really necessary as the defibrillator would normally be switched on during operations at which the heart is exposed and if fibrillation occurs during any other surgical procedure our instrument will be fully ready by the time surgical access to the heart has been achieved.

The various additional facilities which make our unit look complex do not touch the basic simplicity of the pulse giving circuit. It is interesting to note that these extra facilities have been approved and in some cases suggested by the clinical staff who use the defibrillator. This particularly applies to the measurement of the magnitude of the current pulse. Although the value of the resistance of the human heart given by Mr. Dewhurst (50Ω) may well be a fair average, the value found in practice varies considerably and is a function of the magnitude of the current pulse, of heart size, and of the shape and position of the electrodes. On one hand it has been suggested to us (private communication, J. Bekink, Naamloze Vennootschap tot Keuring van Electrotechnische Materialen, Utrechtseweg 310, Arnhem) that due to the high resistance of the heart our value of applied potential (120V r.m.s.) was too *low* while in practice our clinicians have passed between 3 and 4A through the adult human heart, with our defiribillator.

If there are such wide variations of apparent heart resistance our indication of pulse current magnitude would seem to be more of a necessity than a 'dangerous complexity.' The current reading gives immediate indication of the efficacy of the electrode positioning and may thus point at once to a repositioning of the electrodes if the first pulse is not successful. Furthermore, in conjunction with the variable resistor (R_1 our Fig. 1), the initial pulse values can be minimal, the pulse current being increased until successful defibrillation is achieved. This procedure minimizes burn damage and is a particular necessity for younger patients.

We would add that our instrument has been regularly available for nearly two years, and has been used successfully a considerable number of times. A number of similar models have been built for other hospitals and in every case the particular clinician concerned has wanted the additional facilities (or complexities) included when given the choice.

Yours faithfully,

J. PERRY,

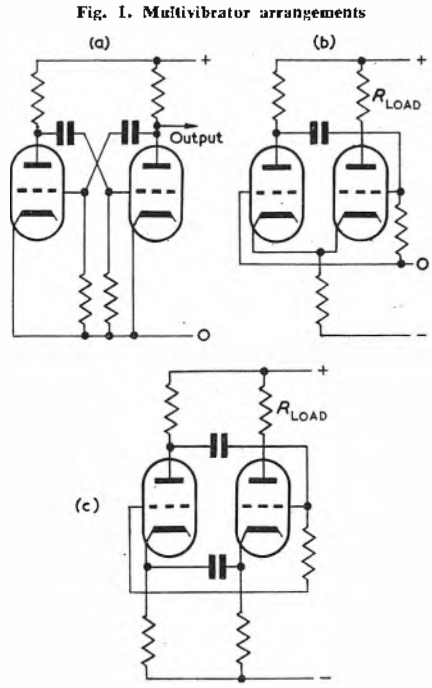
R. E. TROTMAN,

Westminster Hospital,
Physics Department,
Electronics Section,
London, S.W.1.

Multivibrator Circuits

DEAR SIR,—Various multivibrator circuits are described in both reference books and articles in the technical journals. The various forms shown in Fig. 1 are generally described, and the advantages of the 'cathode coupled' forms over the simple circuit of 1(a) are discussed. It is usually stated that the circuits 1(b) and 1(c) improve on 1(a) in having a free anode at which a 'purer' square wave can be obtained than that at either anode of circuit 1(a). Modifications of 1(a), by which a 'squarer' wave may be obtained have been described for use with transistors[1,2].

Fig. 1. Multivibrator arrangements



In none of the references seen by this correspondent has the existence of the 'free' cathodes been recognized in the circuit of Fig. 1(a). The current pulse at these cathodes has a sharp transition at both leading and trailing edges, as shown by the sketches of Fig. 2, which represent the waveforms obtained in the transistor circuits of Fig. 3. The sharp leading spike in 2(c) is probably due to

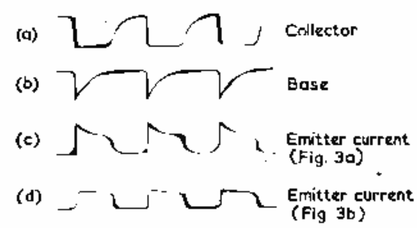


Fig. 2. Multivibrator waveforms

the coupling of the opposite collector pulse through the capacitance $C_{b,e}$ and may be removed very neatly by using a junction diode (such as the OA10) as the emitter load (Fig. 3(b)).

The circuit has been criticized on the ground that the reduced gain (due to degenerative feedback) will unduly lengthen the rise-time of the waveform. This is correct but in a particular multivibrator circuit using OC45's an emitter resistor as large as one-fifth the collector resistor increased the rise-time by a factor of no more than two, both rise times being almost unmeasurable on an oscilloscope with amplifier bandwidth 10Mc/s.

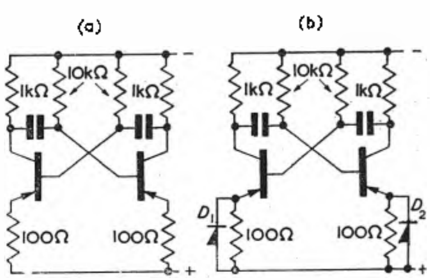


Fig. 3. Transistor multivibrator circuits

The square wave, generated at a level of about 400mV, may easily be amplified if necessary.

Yours faithfully,

M. R. NICHOLLS,

Decca Radar Ltd.

REFERENCES

1. ARMSTRONG, H. L. A Transistor Multivibrator Circuit. (Letter) *Electronic Engng.* 29, 251 (1957).
2. ROZNER, F. Thermistor Multivibrator Circuits. (Letter) *Electronic Engng.* 29, 455 (1957).

BOOK REVIEWS

Feedback Theory and its Applications

By P. H. Hammond. 348 pp. 210 figs. Demy 8vo. The English Universities Press Ltd. 1958. Price 35s.

THIS is the second volume of the new Applied Physics Guides published by The English Universities Press and edited by Sir Graham Sutton. The author is Principal Scientific Officer at the Royal Radar Establishment and the purpose of the book is to present the basic methods of linear and non-linear feedback system analysis at a level suitable for post-graduate engineering and physics students.

After an introduction in which the meaning of feedback is explained, the book begins with a discussion of fundamental concepts and notation. The elements of the Laplace transformation method (described as an operational method) of solving linear differential equations are briefly indicated together with the technique of representing the transformed equations by interconnected block diagrams. The feedback principle and the stability of feedback structures are considered in detail and the discussion includes such topics as Routh's and Nyquist's stability criteria and the response of systems as a function of frequency; the development is in terms of complex-function theory. Some of the foregoing linear theory is then applied to certain electronic networks such as the cathode-follower, the cathode-coupled difference amplifier, and the d.c. coupled virtual earth amplifier which receives special attention because of its importance as an analogue computer element. Finally the linear theory is applied to servomechanisms. The remainder of the book is concerned with the behaviour of non-linear feedback systems. Phase plane and gain describing function analyses are introduced and applied to the study of an electrical position controller servomechanism employing on-off control elements. The final chapter of the book deals with the electronic simulation of control systems.

As the author explains in the preface, many topics necessary for a complete appreciation of modern control system theory have been omitted. Nevertheless the book provides a sound introduction to the subject and will be of immense value to a young graduate about to enter the field of control system engineering.

S. R. DEARDS

Electronic Measuring Instruments

By E. H. W. Barnes. 491 pp. 90 figs. Demy 8vo. 2nd Edition. Chapman & Hall Ltd. 1958. Price 56s.

THE 491 pages of this second edition are divided up into four parts. There are sixteen chapters devoted to very many generalized aspects of electronic measurement.

As is to be expected, and as the author admits in the preface it is not possible to mention all classes of comparable equipment. In fact, it seems a pity that an attempt should have been made to cover so much of electronic measurement under one cover. Useful data that could well have been included has had to be omitted. The result is a book which, upon perusal, turns out to be a flurried mass of brief descriptions about a multiplicity of instruments, none of which seems to give the reader any detailed information.

To the student with no knowledge of electronic measurement such a book might be of assistance in giving some sort of overall picture of equipment likely to be encountered, if he considered it worthwhile to expend the considerable price asked. But owing to the general nature of presentation, as far as the qualified engineer is concerned, little is to be gained, yet at least his knowledge would enable him to ignore badly worded statements and errors in calculations (of which there are some glaring examples) whereas the student might be misled.

There is little or no mention of the huge field of modern communications measuring equipment, which is classified as "test gear". And transistors are dismissed in six pages. V.H.F. and s.h.f. techniques are ignored.

On the other hand, subjects dealt with include indicating instruments, radiation detectors, photoelectric measuring instruments, counters, electro-mechanical transducers and some industrial instruments such as pH meters, moisture meters, electro encephalographs and many others

At times the author pressupposes considerable knowledge on the part of the reader. As a typical example the opening of Chapter III states—"The general principles of the cathode-ray tube will be familiar to the reader." This is an unjustified assumption.

The book is not at all well written, too many of the technical descriptions are irritatingly ambiguous. In short, to be useful, it would seem to point to the reader having at his disposal a well-stocked technical library to supplement the deficiencies that such generalizing renders inevitable.

R. H. MAPPLEBECK

An Approach to Audio Frequency Amplifier Design

126 pp. 90 figs. Demy 8vo. The General Electric Co. Ltd. 1958. Price 10s. 6d.

THIS book, which includes a section on preamplifiers, is expected to appeal particularly to audio designers in industry and to P.A. engineers. It is believed to be the only publication dealing specifically with circuit details of this nature, and it describes designs for a complete range between 5W and 1 100W. The book is available through booksellers.

A Management Guide to Electronic Computers

By W. D. Bell. 390 pp. 60 figs. Demy 8vo. McGraw-Hill Book Co., New York and London. 1957. Price 49s.

THIS book, like many others that have recently been published, explains the use of data processing machines in business and industry.

It differs, however, in three respects from any previous publication, by not going into it in quite so much detail, by giving a series of case histories of the preliminary studies, transition and final installation of machines actually in use at the moment; and finally a chapter which the reviewer found of great interest entitled 'A Look into our Crystal Ball', wherein is described the running of the 'Futura Company'.

In this chapter the author invokes such interesting gadgets as a keyboard and typewriter linked to the central computer in every senior executive's office, where statistics on the company's progress are instantaneously available because the machine is continuously computing the changing position from information supplied via numerous intake points dotted around the works area. He describes how this computer should assist in a rapid re-scheduling when a major hold-up occurs in the plant due to a breakdown in one of the processes.

He quite rightly devotes the longest chapter to explaining programming, and though in the reviewer's opinion he chose an unnecessarily complex example, he shows that computing autocorrelation coefficients is child's play to a computer.

Because the reader's interest will be stimulated he will be left with the feeling of wanting more details than the book provides, nevertheless the author has fulfilled the obligation he undertook when he chose the title of 'A Management Guide to Electronic Computers'.

W. WOODS-HILL

The Ultra High Frequency Performance of Receiving Valves

By W. E. Benham and I. A. Harris. 173 pp. 50 figs. Demy 8vo. Macdonald & Co. (Publishers) Ltd. 1957. Price 28s.

THE authors state that the objects of this book are twofold. Firstly, to present the theory of the conventional valve as a circuit element, enabling linear circuit analysis to be applied to the problem of valve amplifiers for small signals at ultra high frequencies. Secondly, to give a simple, detailed account of the electronic processes in a valve. This account is not only intended to form a basis for the u.h.f. theory of the valve, it is also intended to provide a sound introduction to the transit time theory of thermionic devices and to provide the mathematical apparatus for the investigation of new problems within the scope of the theory. The standard of mathematics demanded of the reader is well within the normal attainments of engineers and physicists. An elementary knowledge of valves, amplifiers, mixers and circuit analysis is assumed.

The Ionosphere: its Significance for Geophysics and Radio Communications

By Karl Rawer. Translated from the German by Ludwig Katz. 202 pp. 71 figs. Demy 8vo. Crosby Lockwood & Son Ltd. 1958. Price 42s.

THE ionosphere as an electrically conducting medium in the upper atmosphere has been the subject of intensive study for the past thirty five years or so, and the amount of literature published in scientific and technical journals is vast. But although nowadays the bulk of the world's communication services are carried out by radio waves which travel between the earth and the ionosphere, the number of books devoted to the subject is lamentably small. The publication of an English translation of this book is therefore to be welcomed, even though the original German edition has been available for over five years.

The author, who was formerly with the French naval ionospheric prediction service, first treats the ionosphere as a particular region of the earth's atmosphere which can be explored by means of radio wave soundings. The first two chapters deal with basic principles of such soundings, with typical observational results and with the manner in which these can be supplemented by a knowledge of aurorae and related geophysical phenomena. Chapter III then proceeds to consider the formation of specific layers in the ionosphere under the influence of solar radiation and the recombination processes which vary with the density of the air.

As a result of regular soundings made at a number of observatories, the general characteristics of the ionosphere are well known, and they are described as regular phenomena in a chapter which deals also with the irregularities associated with magnetic storms and other disturbances. A final chapter deals with the influence of the ionosphere on the propagation of radio waves, and illustrates the manner in which the optimum frequencies for communication over various distances are derived. Brief reference is also made here to the world charts used by various national organizations for predicting such frequencies.

The book is of a suitable length for radio engineers and scientists, and also for geophysicists who require a reasonably straightforward introduction to the ionosphere and its chief characteristics. Although the English is perhaps not always so good as would be expected if it were the author's native tongue, the presentation of the text is, in general, quite clear, and only an elementary knowledge of mathematics is required by the reader. For those who desire to pursue the subject in more detail, a reference list of some two hundred papers is given in the bibliography appended to the book. Its publication at the present time is particularly opportune in view of the extensive investigations of the earth and its atmosphere being made in connexion with the International Geophysical Year.

R. L. SMITH-ROSE.

Problems in Electronics with Solutions

By F. A. Benson. 219 pp. Demy 8vo. E. & F. N. Spon Ltd. 1958. Price 36s.

THIS book has been compiled in order to provide lecturers in the electrical engineering departments of universities and technical colleges with a convenient reference source for problems based on the main sub-divisions in a university undergraduate electronics course. The 282 problems in the book are divided into 21 sections. Wherever possible these problems are based on practical data so as to encourage the student to whom they are set, or who applies himself to them, to think constructively and to develop a sense of magnitudes. Only the electrical steps, and not the step by step mathematics, have been included in the solutions. Diagrams have been freely used wherever they contribute to the better understanding of a solution.

Instrument Encyclopædia

Compiled by E. W. Battey. 292 pp. Demy 4to. The Herbert Publishing Co. Ltd. 1958. Price 63s.

THE purpose of this encyclopædia is to provide a concise reference to the materials and equipment used and produced in the instrument industry, and those industries closely allied to it. It is not intended as a textbook, but is provided as a means of reference. The encyclopædia incorporates a guide to 3 230 classifications of manufactured products which has been divided into the eight sections of present day nomenclatures. The alphabetical index to names and addresses of manufacturers of instruments, components and accessories is amplified by an alphabetical index to agents and distributors and a glossary of trade names used in the instrument industry.

Radio Upkeep and Repairs

By Alfred T. Witts. 234 pp. 174 figs. Sir Isaac Pitman & Sons Ltd. 1957. 8th Edition. Price 15s.

THIS practical handbook explains in an easy to follow style how to locate faults, how to remedy them, and how to keep modern radio receiver apparatus in the best possible working condition.

In this edition new chapters have been added on servicing f.m. receivers and also on servicing transistor circuits and printed circuits.

Frequency Modulation

By A. W. Keen. 274 pp. 80 figs. Sir Isaac Pitman & Sons Ltd. 1958. Price 30s.

THIS book provides for those who are interested in the technicalities of the subject a concise but comprehensive account of both the transmitting and receiving sides of f.m. broadcasting technique. The treatment is entirely descriptive and should be easy to assimilate for anyone having a sound knowledge of existing amplitude modulation technique. The author is Senior Lecturer in Electronics and Telecommunications at Coventry Technical College.

ELECTRONIC EQUIPMENT

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

T.F.A. REFERENCE RESOLVER

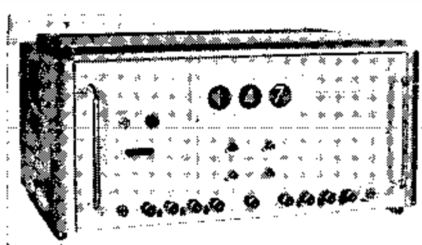
(Illustrated below)

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

THIS instrument, which now has a neon digital presentation, has been designed to give simultaneous identical phase shifts to each phase of a four-phase reference signal. It makes use of Solartron standard printed circuit computer amplifiers. Slight gain adjustment is provided in each of the input channels to adjust for variations in amplitude of the input vectors. A smooth 0 to 360° phase-shift control directly calibrated permits easy setting up of any phase shift required.

When used with the 'Solartron' transfer function analyser an $R\,\theta$ type presentation can be obtained by adjusting the phase shift control to null the quadrature meter. The reference meter then measures R and the phase shift control indicates θ.

The use of the reference resolver also permits allowance to be made for unknown phase shifts in transducers driving the work, without the necessity of plotting two diagrams and computing the difference. The method is to observe the output from the transducer and resolve the reference to be in phase with this output. The Nyquist diagram for the ensuing servo can then be directly plotted in the usual manner.



A further application is the synthesis of vectors for backing off other signals.

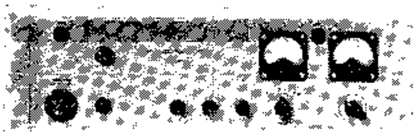
TRANSISTOR STABILIZED POWER PACKS

(Illustrated above right)

Ultra Electric Ltd, Western Avenue, Acton, London, W.3

ULTRA Electric Ltd have developed a series of stabilized power packs using transistors. The units are mains operated and can be mounted if desired on a 19in rack. Output current and voltage can be monitored by panel meters. Model BP.115, for example, can supply up to 2A at any voltage between 2 and 12V. Voltage is adjusted by setting a range switch and a fine control.

The output resistance is less than 0·1Ω over the range and the ripple less than 1mV. For a change of mains input of ±20V the output voltage does not change by more than ±10mV.



A junction rectifier bridge is used to supply current to the load via the stabilizing circuit, which uses transistors. A Zener diode is used as a reference source.

These units are particularly suitable for supplying transistor circuits, for example: A computer using surface barrier transistors where a heavy current stabilized supply at 3V is necessary, power units providing higher currents, higher voltages and for special pulse applications such as feeding ferrite core matrices are in development.

1kW, V.H.F. TRANSMITTER

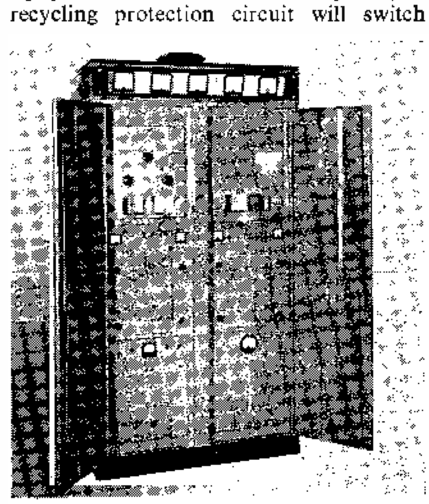
(Illustrated below)

Pye Telecommunications Ltd, Newmarket Road, Cambridge

THE Pye PTC 3600 1kW v.h.f. transmitter is a medium power communications equipment which has been designed primarily for airport ground-to-air operations and also for teleprinter and v.f. point-to-point links. The frequency range from 118 to 136Mc/s is continuously covered. It is basically a two-unit transmitter and the r.f. and modulator sections are assembled in self-contained cubicles which are combined into a composite equipment for r.t. service.

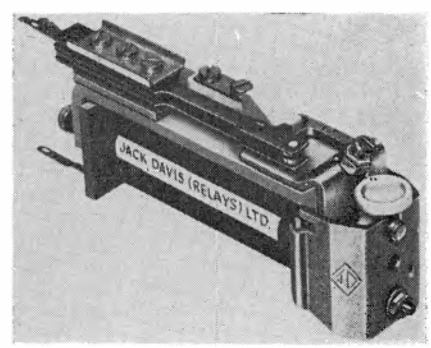
A feature of the transmitter is the ease and rapidity with which it is possible to set it up on any frequency in the range. Due to this, one transmitter can be used as a standby for several others.

The transmitter may be remotely controlled, and is fully protected against damage due to overloads or maladjustments. In the event of a fault in the equipment or in the aerial system, a recycling protection circuit will switch the transmitter on up to three times in succession, and if the fault persists after the third time the transmitter will remain off until the fault is rectified.



Behind the two front doors of the cabinet are 19in panels on which all the operating controls are fitted. Preset controls are concealed by removable front panels. All valves, relays and fuses are accessible from the front of the unit; doors are provided at the rear to facilitate inspection and maintenance.

Valves and components have been chosen and rated to achieve a high standard of reliability and, as a result, the transmitter may be operated unattended in tropical climates. Forced air cooled tetrodes are used in the r.f. output stage and radiation-cooled triodes in the modulator. Filtered cooling air is circulated through the cabinets and is discharged by a fan at the top. All chassis are cadmium plated; transformers and chokes are vacuum impregnated.



RELAY LATCH

(Illustrated above)

Jack Davis (Relays) Ltd, Tudor Place, London, W.1

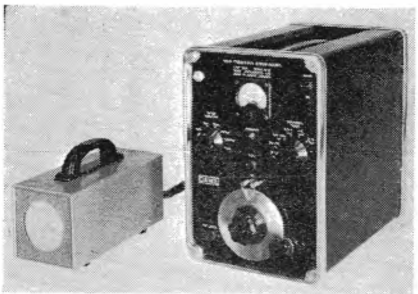
This is a simple device, that can be fitted in about 30 seconds to any 3000 type magnetic relay, for latching the relay after energizing. Release is effected by push-button above the latch or, for remote releasing, by an extended cable.

HIGH FREQUENCY STROBOSCOPE

(Illustrated above right)

Dawe Instruments Ltd, 99 Uxbridge Road, London, W.5

MANY modern machines, ranging from internal grinding spindles to airborne compressors, run at speeds of the order of 100 000rev/min. The normal stroboscopic flash tube is incapable of reaching similar flash rates due to the relatively long de-ionization time of the gas. The operation of such mechanisms is therefore usually studied by flashing only once in every five or six



revolutions. Even at frequencies as low as 250c/s this practice involves a loss of precision and sharpness which becomes even more marked when approaching the higher audio frequencies of 8 or 10kc/s.

The type 1205 high-frequency stroboscope avoids these difficulties of a long de-ionizing time by using as the light source a 2in cathode-ray tube. Flash rates of up to 600 000 per minute (10kc/s) are now achieved by using the variable-frequency square wave from a built-in oscillator to trigger a pulse generator which in turn modulates the intensity of the electron beam of the cathode-ray tube. The pulse duration is variable from 3μsec to 1μsec in six steps, an electrical interlock being provided to prevent overloading of the cathode-ray tube should a longer pulse be accidentally selected at the higher oscillator frequencies.

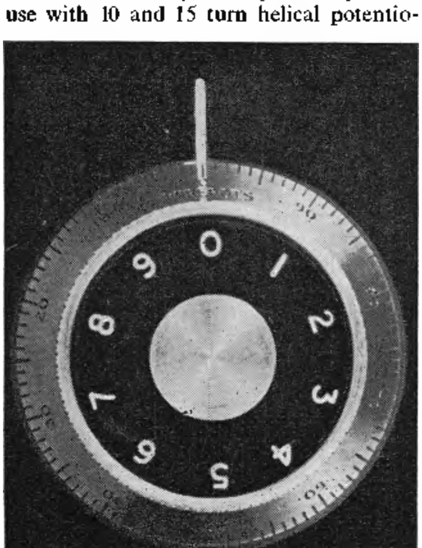
The light output of the unit depends on which of the different coloured phosphors is selected. The standard tube has a phosphor with a green luminescence which gives a light output of some 3 000 candelas and a decay time of less than 1μsec. This output is of the same order as that of a typical neon stroboscopic lamp. Where greater intensity is desirable, a yellow-green phosphor gives an increased light output of around 7 000 candelas with a decay time of 6μsec.

INDICATING DIAL

(Illustrated below)

Wirepots Ltd, New Road, Rainham, Essex

THIS precision and compact indicating dial has been primarily developed for use with 10 and 15 turn helical potentiometers, but it is possible to obtain modified versions for other applications, these being entirely dependent upon the suitability of the internal gearing.



The knob and skirting are machined in one piece from solid duralumin, with the skirting and revolution counter being precision machine engraved. The finger control section is machine knurled.

It is direct gear driven with negligible backlash. Fixing of knob to spindle is by means of an Allen socket head grub screw.

It is supplied as standard in duralumin self colour, with black graduations on the skirting and white numbering on the revolution counter. Various anodized colours with silver graduations and skirting numbers, to blend with existing colour schemes, can be supplied on request.

The overall diameter of the dial is 1¾in and it can be fitted with an anti-vibration locking device. It is also being produced in semi-miniature form.

'RUGGED' E.H.T. RECTIFIER

General Electric Co. Ltd, Magnet House, Kingsway, London, W.C.2

PROBABLY the first 'fully-reliable' e.h.t. rectifier in the world has been produced by the M.O. Valve Co. Ltd. The valve has the title U60 or CV4075.

Owing to the use of a special cathode of large emissive area, a peak cathode current of 300mA is obtainable. P.I.V. is 30kV at 4mA, and the valve is extremely economical in cathode power. It is therefore particularly useful for fly-back e.h.t. circuit applications, especially in radar equipment undergoing vibration.

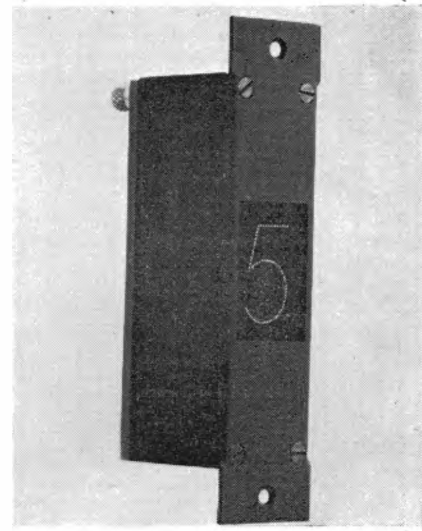
IN-LINE INDICATORS

(Illustrated above right)

K.G.M. Electronics Ltd, The Barons, St. Margaret's, Twickenham, Middlesex

THE K.G.M. multi-indicator has been designed to provide selective display of any of a set of numbers, letters, etc., in the same space as shown in the photograph. This indicator is based on the principle of the edge lighting effect of engraved perspex plates. These plates are stacked together each one of which is 'edge illuminated' by a 'pea' lamp.

A common disadvantage of this type of indicator is that the angle of viewing is severely limited by the overall thickness of the stack of plates, which is necessarily great because of the comparatively large dimension of the lamps. This disadvantage has, however, been overcome in the K.G.M. indicator by utilizing the total internal reflexion characteristic of bent perspex sheet; the light from the illuminated edge is conveyed round the bend without transmission through the surface of the plate. The plates are bent over at right-angles at one end and the tails thus formed are inserted into individual slots behind which are the light sources. This construction permits the plates to be tightly packed together forming a stack only slightly more than ¼in thick and thereby



permitting the rear plate to be viewed at a very wide angle. A further advantage of this construction is that the lamps are easily accessible from the rear. The K.G.M. multi-indicator is designed for fitting into panels of any thickness either singly or in banks and its size has been kept as small as possible commensurate with a suitable size of display. The engraved characters are readable at a distance of twenty feet or more depending upon room illumination.

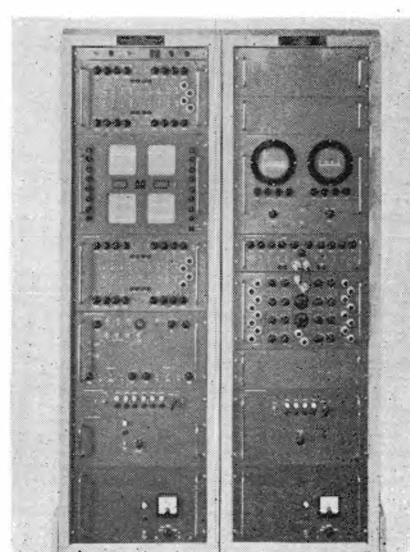
MULTI-CHANNEL RECORDING AND STORAGE OSCILLOSCOPE

(Illustrated below)

Cawkell Research and Electronic Ltd, 6-8 Victory Arcade, Southall, Middlesex

A MULTI-CHANNEL recording and memory storage oscilloscope of a type very similar to that supplied to Harwell for measuring purposes on the ZETA project (Electronic Engineering, p. 15, March, 1958) is now available commercially.

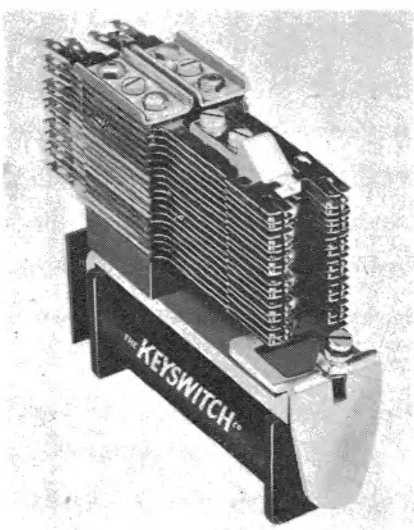
The instrument is believed to be one of the most comprehensive of its type manufactured in this country. The eight-



channel oscilloscope, with provision for photography, and the two-channel storage oscilloscope, are housed in a two bay 6ft rack, which also contains all the associated electronic equipment.

The storage tubes retain the waveform of a non-recurrent transient indefinitely for detailed visual examination, while the waveforms on the four-gun cathode-ray tubes are photographed for subsequent measurements.

The equipment has applications wherever many rapidly changing phenomena require to be acurately measured simultaneously.



MULTI-CONTACT RELAY

(Illustrated above)

The Keyswitch Company, 126 Kensal Road, London, W.10

THIS twelve changeover contact relay (based on the G.P.O. type 3 000 relay) is now in quantity production. It can be supplied for operation on either a.c. or d.c.

AUTOMATIC TEMPERATURE SCANNING EQUIPMENT

(Illustrated in next column)

Fielden Electronics Ltd, Wythenshawe, Manchester

IN many processes it is desirable to keep a check on a large number of temperature points, but in the majority of continuous processes the indication serves no useful purpose so long as the temperatures are at the desired values. To have one instrument on each point would be extravagant both in the cost of the instruments themselves and in the space required to house them. It is preferable therefore, for one equipment to be used to monitor all points and this is the function of the scanning equipment described below.

The temperature is indicated on a large dial approximately 2ft diameter and around the periphery of this dial are the point numbers e.g. 1-50. A motorized scanning switch enables the equipment to examine each point in turn and a clear perspex pointer, tipped black, indicates the number of the point under examination at any one time.

It cannot be expected that the desired value for all points will be the same and with this equipment it is possible to have a different set point for each of the 50 variables. The value of the set point is shown by the position of a red pointer and this can be adjusted by a small potentiometer under the indicator bezel and in line with the temperature point number.

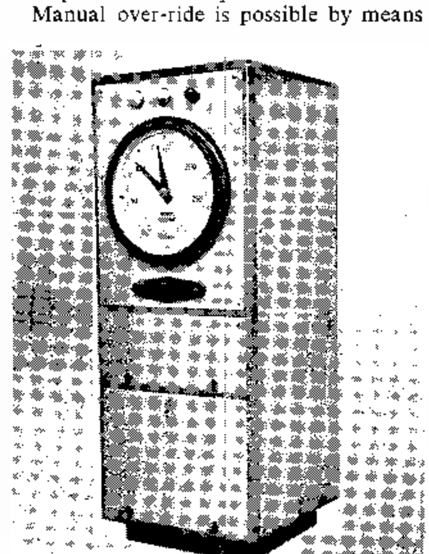
The actual value of the temperature is shown by the position of a green pointer and the measuring circuit is potentiometric. Thermocouples are used as the detecting elements and the output voltage of the thermocouples is compared with a stable d.c. voltage by means of a self-balancing servomechanism. All the thermocouple leads are terminated on a panel at the rear of the equipment and automatic cold junction compensation is incorporated.

It is appreciated that on a large plant certain of the temperature variables may have to be held to closer tolerance than others and to cope with such a situation the equipment has three tolerance bands which can be adjusted to required values. Any number of temperature points can be allocated to any of these 3 bands. This will then enable the more important variables to give an alarm should they deviate by more than, say, 2 per cent, others may be set to give an alarm should they deviate by more than, say, 5 per cent and the least important to give an alarm on a deviation of perhaps 10 per cent.

In normal operation the equipment will scan the 50 points in sequence, the red pointer will balance to the desired value and the green pointer will balance to the actual value. If the difference between these pointers is more than the pre-fixed amount for that particular point, then the scanning switch will stop and the signal lamp corresponding to the particular tolerance band will be illuminated. The scanning switch can be re-started either by bringing the temperature to within the pre-fixed tolerance band or by manually switching the scanner to the next point in the sequence.

Manual over-ride is possible by means of a handle provided on the front panel of the instrument and this enables any temperature point to be indicated at will without waiting for the scanner to reach that point in its normal sequence. Push-buttons are also provided which will stop and start the scanning switch so that this can be arrested on a particular point for any desired length of time.

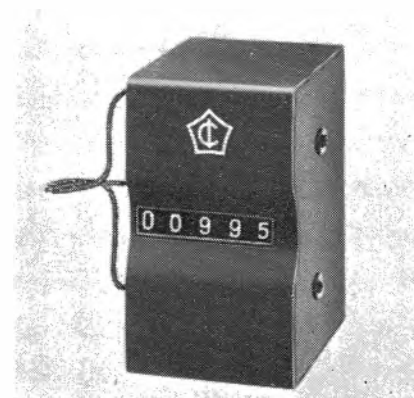
The speed of scanning is approximately 6sec per point and one point can be wired from the constant voltage source in order to check the operation of the equipment.



ADD AND SUBTRACT COUNTER

(Illustrated below)

Counting Instruments Ltd, 5 Elstree Way, Boreham Wood, Hertfordshire.



THIS counter is a self-contained unit with a five figure presentation. The number wheels are actuated by two escapement type mechanisms operated from separate d.c. or a.c. supplies.

The first escapement operates to add, at the same time releasing the second, and when subtracting this operation is reversed.

Counting + or − is possible up to speeds of 17 per second and the design is such that damage cannot occur if both adding and subtracting pulses are supplied at the same time.

Large figures are used in the display and all working parts are made to precise limits to ensure accuracy of count and long life.

The counter is available for operation on 24-28V or 200-250V.

SUB-MINIATURE COAXIAL CABLE

British Insulated Callenders Cables Ltd, 21 Bloomsbury Street, London, W.C.1

THE introduction of a new range of very small coaxial radio frequency cables with overall diameters of less than 0·1in is announced. A typical type has an inner conductor of 1/0·214in silver-plated copper-covered steel wire insulated with solid Polythene, plain copper wire braided and Nylon sheathed to an overall nominal diameter of 0·068in. An alternative sheath of black high density Polythene is also available. The cable has a characteristic impedance of 50Ω and a capacitance at 1kc/s of 30pF/ft. Cables of similar construction but having a characteristic impedance of 75Ω will also be available shortly.