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Commentary

ASTRONOMY is, perhaps, the oldest science known to man; radio astronomy is probably the newest. In the eighteenth and nineteenth centuries, when the Earl of Rosse and Sir William Herschel pioneered the construction of large telescopes, British astronomy achieved considerable distinction but, due to climatic conditions, this country's leadership could not prevail and the centres of astronomical discovery moved to countries with more kindly climates. During the present century the main centres of discovery have been in America, where several very powerful optical telescopes have been built, the most notable being that at Mount Polomar. With the advent of radio astronomy with its independence from cloud and fog, Britain can compete in the fascinating exploration of outer space without handicap. And this she is doing wholeheartedly and in no mean measure.

Although it was known before the war that radio waves reached the earth from outer space experimental techniques at that time were not sufficiently good for their properties to be studied in detail. During the last ten years, however, methods developed for use in radar have been employed in the study of these waves and, out of this study, the important new science of radio astronomy has grown. The object of this science is to discover as much as possible about these radiations and in particular to find out where they come from, how they are emitted and what light they shed upon the many unsolved problems of astronomy.

The research effort in this country is based at the Universities of Manchester and Cambridge. At Manchester, where the team is led by Professor A. C. B. Lovell, the world's largest steerable radio telescope is in an advanced stage of construction at Jodrell Bank; while in the Cavendish Laboratory at Cambridge, Mr. Martin Ryle, F.R.S., and his helpers have made considerable progress using 'interferometer' techniques. Using the latter methods it is possible to provide much of the information that is required for radio astronomy using comparatively simple aerials and equipment.

The success of the first large interferometer, which was constructed in 1952, showed that a still larger one could be expected to yield even more information. Due largely to a generous grant from Mullard Ltd it has been possible to rent a site at Lord's Bridge, about five miles from Cambridge, and to construct there a still larger interferometer for the study of radio stars, and a second one for the study of waves coming from the Milky Way. The Mullard Radio Astronomy Observatory was formally opened by Sir Edward Appleton at the end of July and, as N. F. Mott, the Cavendish Professor of Experimental Physics, pointed out at the opening ceremony, the gift of £100 000 by Mullards has made it possible to plan the project as it should be planned and in fact represents more than half the total expenditure. A grant has also been made by the Department of Scientific and Industrial Research and this venture represents an outstanding example of the co-operation between industry, government and university for the furtherance of scientific research in this country.

The interferometer techniques used at Cambridge are in contrast, and are complementary to, those used by Professor Lovell at Jodrell Bank where the large and complex steerable telescope is under construction. The two pieces of apparatus will be used for the investigation of the different problems for which they are best suited and, although there will be a certain amount of friendly rivalry between the two teams, they will co-operate in solving the many problems of radio astronomy; they will also co-operate with the 'optical' astronomers at Mount Polomar. The interferometer is being built primarily for a specific wavelength, which it will not be easy to change. This coupled with the fact that the instrument, although steerable in declination, will not be able to track, makes possible a simpler and more crude structure at considerably less cost. It will be used to study the statistical distribution and to obtain precise positional measurements of the large number of radio sources. It will also be used to investigate galactic structure on wavelengths of the order of 8m. The Jodrell Bank telescope, on the other hand, will be more concerned with obtaining detailed information about a particular radio source; for example, the diameter and spectra of the source, together with background surveys on wavelengths shorter than is possible with the Cambridge interferometer. The facility with which the wavelength can be changed on the radio telescope makes it particularly suitable for these purposes. In addition, it has a wide range of uses in the radio echo aspects of radio astronomy which is outside the scope of the interferometer.

When both instruments are working this country will have two very able research teams each supplied with unusually powerful equipment and we will be set fair to make our bid to become the undisputed leaders in this modern form of a very old science.

Circuit Techniques Associated with

Transistor Broadcast Receivers

(Part 1)

By J. N. Barry, M.Sc.

This article outlines some of the special circuit problems which arise due to the use of transistors in sound broadcast receivers, and discusses technical solutions which can be adopted to overcome them. The particular cases of battery portable and car receivers are both considered in some detail. An assessment of the effect of these technical solutions on receiver performance and economics is also made, and detailed performance figures for six different receiver designs are presented. The latter include four transistor portables, a valve portable and a transistor car radio. In the concluding section, any differences which the introduction of more recent types of transistor might make are discussed briefly, with particular reference to the case of v.h.f., f.m. receivers.

BROADCAST receivers may be divided broadly into three categories, i.e.

(a) Mains driven.
(b) Battery operated.
(c) Car radios.

The use of transistors in a normal mains driven receiver offers, at the present time, no significant technical advantage; in fact it would suffer from economic disadvantages compared with its valve counterpart due to the relatively high cost of transistors, and the fact that a larger number would be required to give an equivalent electrical performance. In the other two categories, while the initial manufacturing costs might still not compare favourably with the valve equivalent (particularly in the case of the simple battery receiver), it is apparent that the use of transistors offers certain definite attractions, including small size, light weight and substantial operating economies.

The considerations presented in this article apply in general to either of the latter two cases, though special requirements of either case are treated separately.

As far as the broadcasting field is concerned it is apparent that it can be classified broadly into two systems, i.e. a.m. and v.h.f., f.m. As the useful frequency limit of currently available transistors in the United Kingdom is of the order of 10Mc/s, while the frequency range of the v.h.f. band is 88Mc/s to 100Mc/s, there appears to be little likelihood of designing suitable commercial all-transistor receivers for this band at the present time, and hence most of the problems discussed deal specifically with circuits intended for operation in the lower frequency a.m. bands.

Throughout the article the use of junction type transistors is implied. Although the use of point contact transistors is possible for certain functions in a broadcast receiver, e.g. as local oscillators, their use as amplifiers is not attractive. This is due to the fact that although the frequency range of point contact transistors originally exceeded that of junction types, their high noise factors, tendency to circuit instability, and their high current consumption relative to junction transistors rendered them unattractive for receiver designs. In considering current circuit performance the frequency limitation of available transistors has been borne in mind. In the concluding section of the article, however, some of the most recent experimental developments in junction transistors are considered, and their possible application in the broadcast receiver field, including the short-wave a.m. and v.h.f., f.m. band, is discussed briefly.

As the logical development of transistor receivers has taken place from the audio stages through to the r.f. circuits, the problems arising are considered in this order.

The Stages of the Receiver

General Considerations

In considering the design of circuits associated with transistors in broadcast receivers a number of problems arise. Most of these are due to certain fundamental differences between the operation of transistors and valves. Thus compared with valves transistors are relatively low impedance devices; they therefore function most usefully as power amplifiers, and power matching between the stages is of considerable importance. Transistors can also have appreciable internal junction capacitance (that between collector and base being of particular importance), and this means that transmission in the reverse direction can be a significant factor, particularly with operation at higher frequencies[1, 2].

Most of the parameters of a transistor are dependent on both temperature and d.c. operating conditions, and the effect of such variations has to be catered for in addition to achieving satisfactory electrical performance. It is shown that special circuit techniques are often needed to minimize these effects. The additional circuit complexity is usually relatively small, but it can tend to increase the adverse economic balance of all-transistor receivers.

Two other important considerations concerning transistor amplifiers are their noise and frequency performance. Although the noise factors of earlier transistors were appreciably inferior to those of valves, more recent types have shown a substantial improvement in this respect. Minimum noise factors of present junction types can be of the order of 6dB at 1kc/s. For the higher frequency transistors this figure can drop to the order of 2 to 3dB at 500kc/s.

At the present time, excepting for the moment some of the most recent experimental transistors announced in America, the upper limit for amplification with good power gain is only of the order of a few megacycles per second. It is found that the practical stage gain of transistor ampli-

* *Research Laboratories of The General Electric Co. Ltd., Wembley.*

fiers at 470kc/s, a typical standard intermediate frequency, is of the order of 6 to 10dB less than that of an equivalent valve stage. For this reason receiver design tendency to date has been to place as much as possible of the required gain following the second detector, in order to minimize the gain required at high frequency. For similar reasons the use of an r.f. stage preceding the mixer is rather an embarrassment. The omission of this stage is standard practice in the case of portable receivers, but for car receivers, where very good image rejection is required, the use of such a stage is usually considered necessary.

THE AUDIO STAGES

These are almost invariably common emitter operated, and the circuit design is by now fairly well established. One important problem to be considered however is that of thermal stability. This is of importance due to the fact that the maximum value of the temperature dependent reverse current in the common emitter arrangement: I_{ceo}, is given by:

$$I_{ceo} = \frac{1}{1-\alpha_0} I_{cbo}$$

i.e. $I_{ceo} = (\alpha_{cb})_0 I_{cbo}$

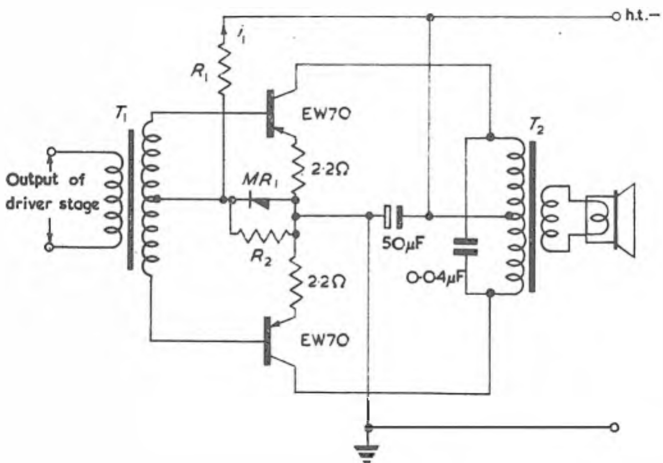

Fig. 1. Push-pull circuit with bias stabilization

where $(\alpha_{cb})_0$ = low frequency current gain, collector-base.
I_{cbo} = collector cut-off current (emitter open-circuit).

The magnitude of I_{ceo} can hence have a value of an appreciable fraction of a milliampere even at room temperature, and can produce a significant rise in junction temperature. It is therefore found that unless the effective d.c. gain of such a stage is strictly controlled thermal instability or 'runaway' can occur, due to the rapid build-up of the current-temperature effect.

This consideration is of importance in both the driver and output stages, where appreciable transistor dissipations are necessary due to signal power requirements. In such cases, unless the d.c. source resistance is kept very low, some d.c. negative feedback is required, this usually taking the form of added resistance in the emitter circuit. In practical circuits the most rigorous conditions from the thermal stability standpoint occur in a class-A power amplifier due to the higher quiescent power dissipation in this case (i.e. when compared to a class-B stage). The method of calculating the amount of feedback required to ensure stability is given in the Appendix.

The audio output stage usually consists of two transistors operated in a class-B push-pull circuit. This is chosen both on account of its very low current consumption under quiescent signal conditions (of particular importance for battery portable receivers), and because of its high operating efficiency.

The forward current bias is restricted to that required to overcome cross-over distortion[3], and is usually of the order of 1 or 2mA for battery portable receivers, but possibly up to 10mA for stages delivering powers of a few watts such as for car radios.

It is usual to stabilize the forward bias against changes of either h.t. voltage or temperature. In addition to securing thermal stability, this is necessary to avoid distortion occurring should the forward bias be varied appreciably. Temperature is of particular importance in car radio applications where a range of 0°C to 45°C might easily be encountered. The circuit shown in Fig. 1 is one method of performing both functions simultaneously, the principle of operation being as follows.

The transistors are forward biased by means of the potentiometer network R_1 and MR_1, the diode operating in its forward or low resistance direction. R_2 is appreciably larger than the forward resistance of MR_1 and is incorporated to avoid a dangerously large forward bias being applied should MR_1 go open-circuit. The forward voltage-current relationship of MR_1 may be written:

$$I = I_S (e^{qV/kT} - 1) \quad \ldots\ldots\ldots\ldots \quad (1)$$

where I_S = the reverse saturation current, and
k = Boltzmann's constant.

Since the forward characteristic where $I \gg I_S$ is being considered, equation (1) approximates to:

$$I \simeq I_S e^{qV/kT} \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

If the resistance R_1 is also much larger than the forward resistance of MR_1, so that the current i_1 (see Fig. 1) may be written:

$$i_1 \simeq (v_{h.t.}/R_1) \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

then it may be shown that for a given ratio of h.t. voltages, the corresponding ratio of forward bias voltages (across MR_1), is given by:

$$\frac{v_{b1}}{v_{b2}} \simeq 1 + \frac{\ln \dfrac{(v_{h.t.})_1}{(v_{h.t.})_2}}{\ln \dfrac{(v_{h.t.})_2}{I_S.R_1}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

where $(v_{h.t.})_1 > (v_{h.t.})_2$.

Due to the non-linear nature of the forward characteristic of the transistors' emitter-base junctions, the forward bias current depends markedly on the forward bias voltage. The second term on the r.h.s. of equation (4) is usually very small, so that good stabilization is achieved.

In addition, if diode MR_1 is of the same construction as the output transistors (e.g. both germanium junction devices), good temperature compensation is achieved as the emitter-base diodes of the transistors are connected back to back with diode MR_1.

Experimental results using a germanium junction diode for MR_1 in the above circuit in a car radio application have shown that the bias current should vary by less than 10 per cent over a temperature range of 5°C to 35°C. Values for R_1 and R_2 of about 200 to 300Ω and 100Ω respectively are usually of the right order for stages delivering powers of one or two watts.

If temperature variations only are of importance it is possible alternatively to use a suitable thermistor[9] in place of MR_1. In a practical case economic considerations will probably determine the particular type of compensation circuit to be used.

In a class-B stage maximum operating efficiency occurs at full output, but the maximum collector dissipation per transistor occurs at approximately half the maximum output. If the collector efficiency has its maximum value of 78·5 per cent, it can be shown that the maximum output power P_{max} is related to the maximum permissible collector dissipation P_c by the relation:

$$P_{max} = (\pi^2/2) \, . \, P_c \simeq 4{\cdot}9 \, P_c$$

In practice, as a small amount of forward bias is used, and as the transistors have a small 'knee' voltage, a figure of $P_{max} = 4 \, P_c$ is usually a better representation of practical conditions. This corresponds to a collector efficiency of just over 70 per cent at full output.

Although the mean power output in a broadcast receiver is usually appreciably below the maximum, it should be remembered that overload conditions initiated by the user must be catered for, in which case the mean power output

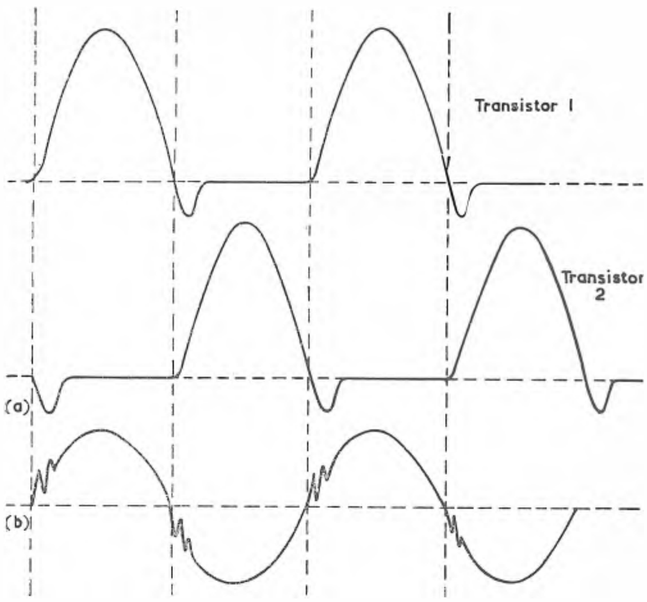

Fig. 2. Waveforms in class-B push-pull output stage

might rise considerably. The relation between P_{max} and P_c at the highest operating temperature envisaged should not therefore be exceeded.

As in radio receiver applications maximum power gain is usually of greater importance than minimum distortion[3], the common emitter amplifier arrangement is usually used. Negative feedback can be applied if required to improve the output waveform. For common emitter stages this is done by connecting a resistive feedback circuit between the output load and the input of the driver, and in this way it is usually possible to keep the total harmonic distortion to less than 10 per cent, unless the output transistors are badly mismatched. The effect of mismatching the output transistors in a simple class-B push-pull stage is treated fairly fully in a previous paper[4].

A particular difficulty associated with class-B push-pull circuits is their tendency to spurious ringing effects unless the circuit components, especially the driver and output transformers, are carefully designed. This tendency is enhanced in the case of transistor circuits due to carrier storage effects i.e. the fact that a finite time is required to remove all the minority carriers (holes in a pnp transistor) from the base layer when the base-emitter junction is biased in the reverse direction. This means that as the operating frequency is raised, it may not be possible to cut-off completely one half of the circuit before the other half starts conducting. The input current waveforms at the two base terminals will hence be as shown in Fig. 2(a), and this will give rise to pulses appearing at the crossover positions in the output waveform. Such pulses will therefore occur at times when both transistors present high impedances to their respective halves of the output transformer primary winding, and can give rise to 'ringing' in the output voltage waveform, as shown in Fig. 2(b). The frequency of the parasitic oscillation is determined by the leakage inductance and self-capacitance of the transformer windings, and is usually of the order of 100kc/s.

For broadcast receiver applications, where the upper audio response is usually limited to about 4kc/s, these undesirable

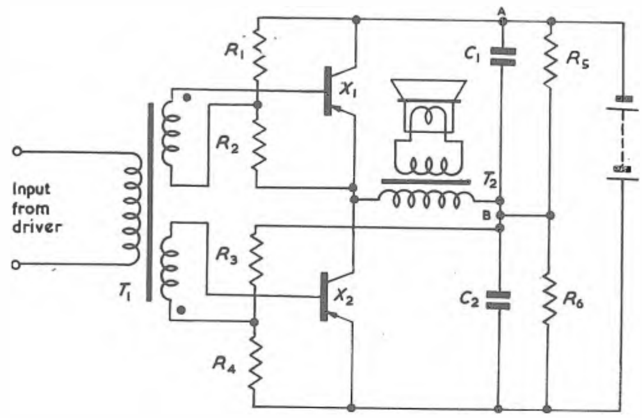

Fig. 3. Modified class-B push-pull stage

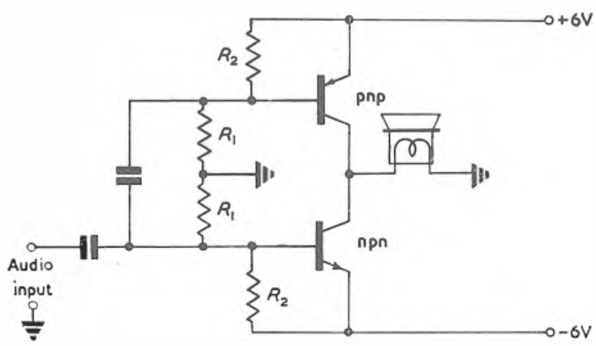

Fig. 4. Basic complementary symmetry common emitter circuit

effects can usually be eliminated by connecting a capacitor of suitable value across the primary of the output transformer.

For high fidelity audio output stages however other precautions might be required. Such measures could include:

(a) Using bilfilar wound input and output transformers.

(b) Using transistors having higher cut-off frequencies.

An alternative circuit arrangement for operating a class-B push-pull amplifier is shown in Fig. 3. It will be seen that with this type of circuit the transistors are effectively in series as far as the d.c. supply is concerned, but supply the a.c. load in parallel. Compared with the more conventional circuit the advantages of this arrangement are firstly, that a single-ended output transformer only is required, which, for high output powers, might even be eliminated altogether, the speaker itself being used as the load. This feature should also aid in reducing the tendency to spurious transients. The second important advantage is that for a given h.t. supply the maximum voltage swing between collector and emitter is only one half of that for the conventional push-pull circuit.

The disadvantages are that, unless a centre-tapped battery is used, two large tank capacitors C_1 and C_2 are needed (of value possibly up to several hundred microfarads), and a relatively large bleed current via R_5 and R_6 is necessary in order to stabilize the potential of the point B. (In practice $R_5 = R_6$ and the potential at B is half the battery voltage.) The transistor biasing circuits also involve additional components. Another consideration is that of the power gain of the stage. For the same h.t. voltage and output power the effective load presented to each transistor is only one quarter of that for the conventional push-pull arrangement, and hence the power gain is likely to be of the order of 6dB less.

The modified circuit is only likely to be of significance in car radio applications, where the additional current drain may be permissible.

Before leaving this section mention should be made of the possible use of complementary symmetry arrangements[5]. These involve the use of a combination of a pnp and an npn transistor arranged as a push-pull pair having a single-ended output and again usually class-B operated. A simple basic circuit for a common emitter amplifier arrangement is shown in Fig. 4. The principle of operation

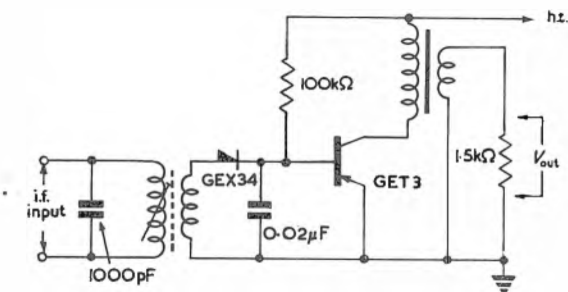

Fig. 5. D.C. coupled diode-transistor detector circuit

of such a circuit is essentially the same as the single-ended output circuit of Fig. 3, and is subject to certain of the disadvantages of the latter arrangement (chiefly special power supply requirements). It has the further disadvantage of introducing additional transistor design problems to ensure good matching, and for these reasons the circuit has not yet been used commercially, as far as is known. It has however the advantage of being single-ended throughout, and therefore does not require any transformers. This factor might become an important consideration if the transistor problems could be solved satisfactorily.

Second Detector Stage

Two choices are possible for this stage, i.e. either the use of a crystal diode in a conventional type of detector circuit, or the use of a transistor to combine the functions of detector and audio amplifier. The choice usually rests with the arrangement capable of providing the better overall detection at low input signal levels. As a diode detector circuit alone entails a conversion loss, whereas the transistor 'detector-amplifier' circuit usually has a conversion gain, the fairest comparison of performance is between the latter and a crystal diode circuit combined with a following audio amplifier stage.

It has been found experimentally that the minimum voltage level at which essentially linear detection can be obtained is very similar for either type of circuit and is of the order of 50mV in value. As the input impedance of the diode circuit is higher than that of the transistor, the minimum permissible input signal power to the detector is therefore somewhat lower in the former case.

In both cases it is necessary to provide a small amount of forward bias in order to minimize distortion at low input signal levels. Distortion can also be reduced by increasing the source resistance, though in the case of a transistor this decreases the conversion efficiency.

As far as conversion gain is concerned the arrangement of a crystal diode detector followed by a common emitter transistor amplifier has a slight advantage, typical values being of the order of 12 to 16dB conversion gain for this circuit compared with 6 to 10dB for a single transistor arrangement, when using a 30 per cent modulated input signal.

However, from the aspect of economy of components, the transistor detector arrangement has some advantages, and neither type of circuit seems to have been given overriding preference in practical receivers.

A particular circuit which has been found successful in practice consists of a crystal diode which is d.c. coupled

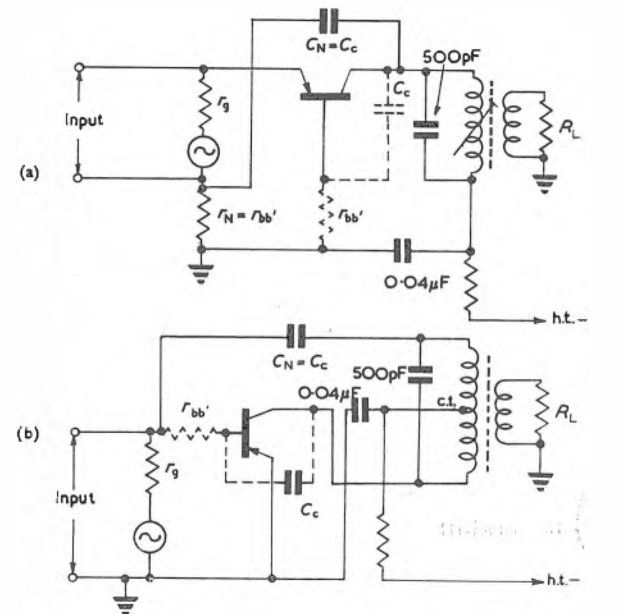

Fig. 6. Neutralized common-base (a) and common-emitter (b) i.f. amplifier circuits

to the base input of a following transistor amplifier as shown in Fig. 5. The advantage of this arrangement lies in the fact that while the performance closely approaches that of the conventional diode plus transistor circuit, it comprises a much more economical arrangement from the component aspect, and provided a suitable d.c. load is used, can still provide adequate a.g.c. power. A particularly interesting feature is that any hole storage effects occurring in the base layer of the transistor, which might produce distortion of the output waveform at high signal levels, are swamped by the detecting action of the preceding crystal diode. On the other hand this circuit produces rather more distortion than either the diode or transistor alone at low input signal levels.

I.F. Amplifier Stages

While the basic design of those stages of a receiver which follow the second detector has been fairly well established, the circuits associated with the i.f. amplifier have been much more flexible. This is primarily due to the continual development process to which junction transistors have been subject in recent years. Before describing any specific i.f. amplifier circuits, it is useful to discuss the principal factors affecting the high frequency performance of junction transistors in small signal amplifiers.

Two performance criteria of extreme importance to the circuit designer are the maximum power gain and the conditions of stability. The latter are dependent on the internal feedback conditions in the transistor, i.e., the reverse transmission characteristic of the representative equivalent circuit network. It can be shown that such feedback from the output to the input circuit takes place via the internal collector-base capacitance of the transistor, and can in some cases (particularly when using common emitter circuits) be regenerative in nature, thus giving rise to instability troubles. Such difficulties are encountered in practice, being especially prevalent when using cascaded stages. It is hence usually desirable to remove or 'neutralize' such feedback effects, and this can be achieved fairly readily by simple circuit techniques. The circuits shown in Figs. 6(a) and 6(b) show examples of neutralized i.f. amplifiers using the common base and common emitter arrangements respectively. When the components C_N, r_N of the neutralizing circuit of Fig. 6(a) are given the values indicated, where

C_c = the collector-base capacitance,

r_{bb}' = the 'ohmic' or 'extrinsic' base resistance,

exact neutralization results. This technique is often referred to as 'Bridge Neutralization' as the complete circuit can be re-drawn as a bridge when it is easily seen that the effect of the two feedback paths cancel (n.b. the internal feedback is shown dotted in both circuits of Fig. 6, and the transistor's extrinsic base resistance r_{bb}' has been drawn as a circuit element to show its influence on performance).

If the added feedback capacitor in Fig. 6(b) is made equal to C_c, a close approximation to exact neutralization results. The principle of this circuit is to introduce an additional feedback component in anti-phase to that existing due to the transistor's internal capacitance. An even closer approach to the neutralized condition can be achieved by adding a series resistor to the feedback capacitor, but this final elaboration is seldom found to be essential in practice, the capacitor exercising the predominant effect.

The effect of omitting the resistor from the neutralizing circuit can be seen from the curves of Fig. 7. The collector circuit was actually tuned to 540kc/s in this experiment, and the transistor used had the following parameters:

$$f_\alpha = 4\text{Mc/s}$$
$$r_{bb}' = 90\Omega$$
$$C_c = 10\text{pF}$$

The effect of internal feedback on the stability of cascaded i.f. amplifier stages has been discussed in some detail in a paper by Holmes and Stanley[6]. For a single stage amplifier, common emitter connected, employing a simple tuned circuit as collector load, they show that the maximum feedback capacitor which can be tolerated between the output and input terminals before instability occurs is given by:

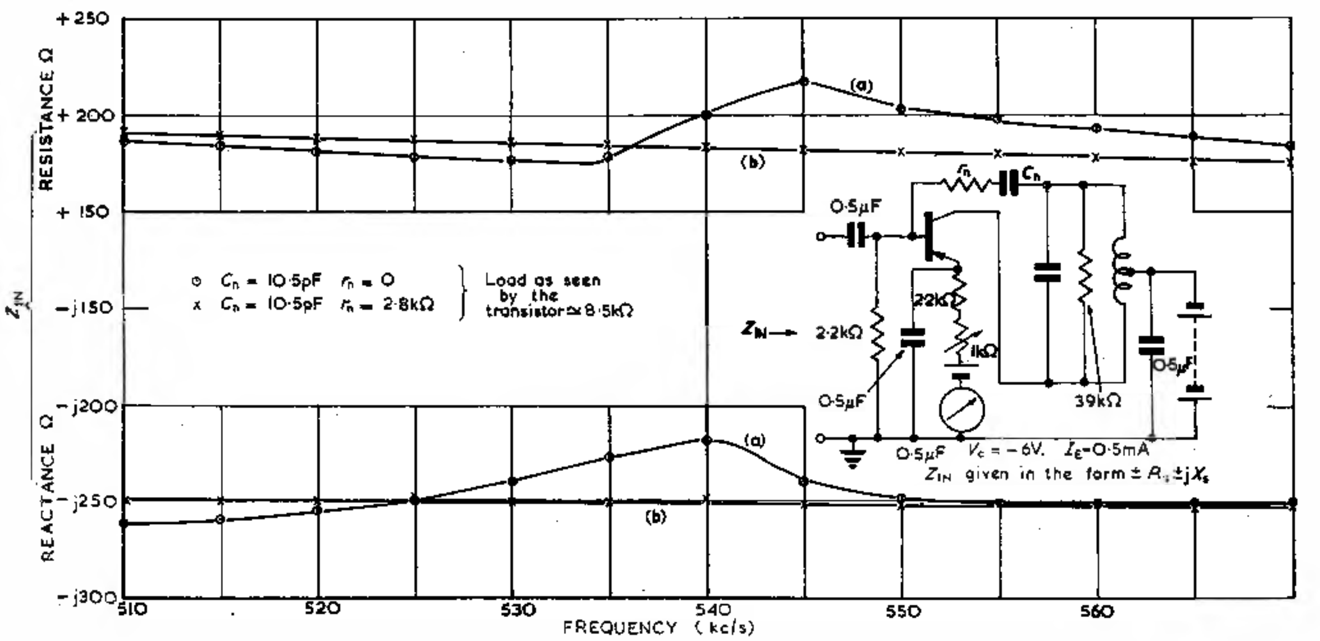

Fig. 7. Common emitter circuit; variation of input impedance with frequency for different neutralizing arrangements

$$C_f = \frac{2}{2\pi f g_m R_1 R_2} \quad \ldots\ldots\ldots\ldots\ (5)$$

where $g_m R_2$ = voltage gain of stage at resonance.

$$R_1 = \frac{R_g R_{IN}}{R_{IN} + R_g} \quad \ldots\ldots\ldots\ldots\ (6)$$

and
$$R_2 = \frac{R_{out} R_L}{R_{out} + R_L} \quad \ldots\ldots\ldots\ldots\ (7)$$

In equations (6) and (7) above:

R_{IN} = input resistance of stage.

R_{out} = output resistance of stage.

For a transistor having an 'α' cut-off frequency of 5Mc/s, and operated at 470kc/s, an approximate value for C_f can be calculated. If it is assumed that:

$$R_g = R_{IN} = 500\Omega$$
$$R_L = R_{out} = 30\text{k}\Omega$$

and $g_m R_2 = \dfrac{\alpha R_L}{r_e + r_b(1-\alpha)} = 400$

then $C_f \leqslant 6{\cdot}5\text{pF}$.

Allowing an arbitrary safety factor of the order of 3 : 1, $C_f \leqslant 2$pF is probably about the limit for reliable operation, and for any value larger than this neutralization would be desirable. When considering cascaded stages the safety factor must be increased still further, and the use of neutralization is even more essential.

As will be seen from equation (5) the maximum permissible value for C_f would be increased if either R_1, R_2 or g_m were decreased. In practice such changes would result in loss of gain. In the above analysis C_f can, to a first approximation, be considered as the maximum permissible collector-base capacitance of the transistor.

in the i.f. selectivity characteristic. The transistor used in obtaining Fig. 8 was the same as that previously specified.

From the preceding discussion it will be evident that, unless steps are taken to counteract the effect of internal feedback, the power gain of a stage will be governed primarily by the feedback conditions existing, and can even become infinite under conditions producing oscillations. It is hence really useful to consider only the maximum available gain under neutralized conditions*.

Maximum available gain infers that impedance matching is achieved at both the input and output terminals; i.e. the generator impedance is the conjugate of the input impedance, and the load is equal to the transistor output resistance (the reactive component of the transistor's output impedance is usually included in the tuning of the collector circuit). In practice the gains obtained are always slightly less than the maximum available due to circuit losses, but by careful transformer design a practical figure which is close to the theoretical can be obtained.

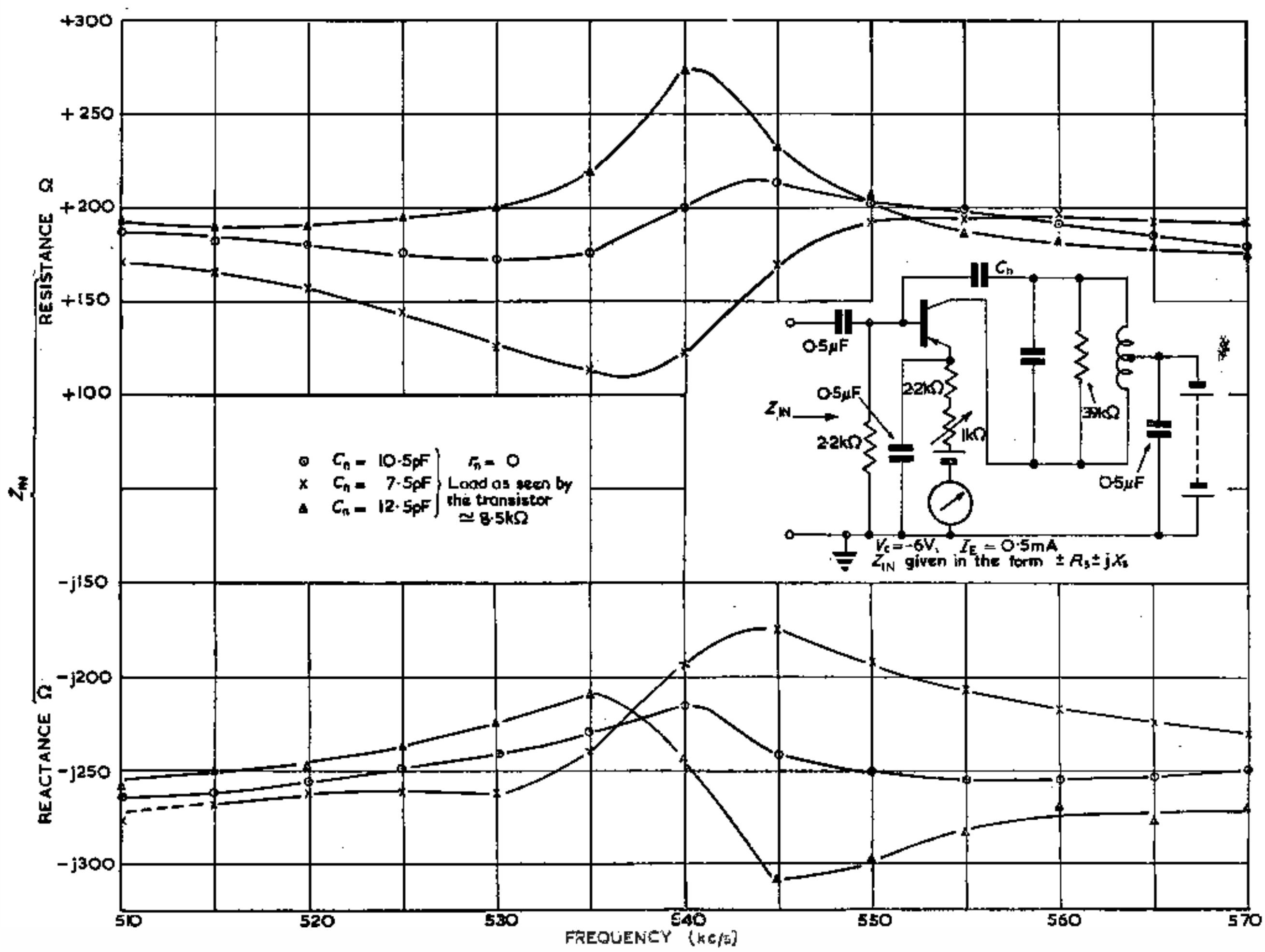

Fig. 8. Common emitter circuit; variation of input impedance with frequency for different values of neutralizing capacitor

The value of C_f given by equation (5) will also give an indication of the tolerance permissible on the neutralizing capacitor for a circuit such as shown in Fig. 6(b). In such cases the tolerance is usually not less than ± 25 per cent of the correct value of capacitance. Variation in the conditions of neutralization may result either from the tolerances on nominal component values, or due to variation in the receiver h.t. voltage, which will cause a variation in the value of C_c. The curves of Fig. 8 show the effect on input impedance of varying the neutralizing capacitor by approximately ± 25 per cent. Such variations are probably fairly acceptable, though they can produce some asymmetry

A number of papers have recently been published in America giving a detailed account of neutralizing techniques and theoretical predictions of maximum power gain[1, 2, 7, 8]. These show that over a limited frequency range near to the 'α' cut-off frequency the maximum gain in either the common base or common emitter arrangement is a function of frequency, and, in terms of the transistor parameters can be expressed as:

* *Unilateralization is a special case of neutralization in which both the resistive and reactive components of the feedback are cancelled, thus turning a bilateral device into a unilateral device.*

$$G_{max} \simeq (1/8\pi) \,.\, (1/f^2) \,.\, \frac{f\alpha}{r_{bb}'C_c} \quad \ldots\ldots\ldots \quad (8)$$

where $f\alpha$ = α cut-off frequency, and r_{bb}', C_c are as defined previously.

Equation (8) is only valid for transistors of the alloy type, which it is assumed can be considered as having a constant lumped ohmic base resistance r_{bb}'. This type of junction transistor is the one most commonly used at the present time. The frequency range over which equation (8) applies is approximately:

$$0{\cdot}1 < (f/f\alpha) < 2 \quad \ldots\ldots\ldots\ldots \quad (9)$$

The form of equation (8) is a result of a study of the frequency dependence of the parameters of an equivalent circuit representing the junction transistor at high frequencies[7, 8].

It may also be shown that at any frequency the maximum gain of the common base amplifier should never exceed that of the common emitter though it can closely approach it. The reason that the common emitter gain is usually quoted as being appreciably higher than that of the common base arrangement at lower frequencies, is that in

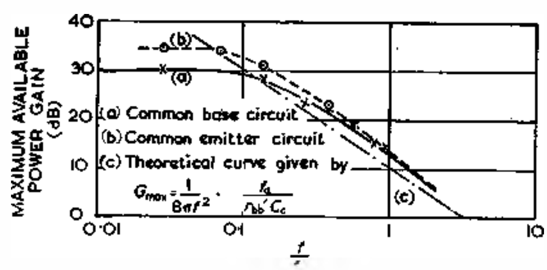

Fig. 9. Maximum gain variation with frequency

order to achieve maximum gain in the latter case very high resistance loads, of the order of 1MΩ, would be required.

The curves (a) and (b) of Fig. 9 show the results of practical measurements of maximum available gain versus frequency for an alloy junction transistor (type 2N139) having an 'α' cut-off frequency of about 4Mc/s. The curves are for common base and common emitter operation respectively, and show the relatively small difference between the two cases. It may be noted that the difference is only some 4·5dB after the curves have flattened out at the lower frequencies. Also plotted on the same diagram is curve (c), which is the value of the maximum gain against frequency as predicted by the theoretical relationship of equation (8). It is apparent that the theoretical curve gives quite good agreement with the measured values down to frequencies of the order of $0{\cdot}1f\alpha$.

For the transistor used, the various parameters had the following measured values:

d.c. conditions: $V_c = -6\text{V}, I_e = 1\text{mA}$

$$\alpha_o = 0{\cdot}97$$
$$f\alpha = 3{\cdot}7\text{Mc/s}$$
$$r_{bb}' = 90\Omega$$
$$C_c = 10\text{pF}.$$

From the above discussion it will be seen that for junction transistors of the alloy type the maximum power gain at any frequency within the range covered by conditions given in equation (9) is governed by the value of the transistor parameter M where:

$$M = (f\alpha / r_{bb}'.C_c) \quad \ldots\ldots\ldots\ldots \quad (10)$$

M is often termed the 'figure of merit'.

It will be noted that in the frequency range considered the maximum gain is independent to a first order of either of the current gain factors, 'α' or 'α_{cb}'.

For grown junction transistors R. L. Pritchard[8] has shown that the corresponding figure of merit is given by:

$$M' = (1/C_c) \,.\, \sqrt{(f\alpha / r_{bb}')} \quad \ldots\ldots\ldots\ldots \quad (11)$$

and that the maximum gain in this case is:

$$G_{max}' \simeq \frac{f\alpha^{\frac{1}{2}}}{10\pi C_c . f^{3/2} (r_{bb}'.r_e)^{\frac{1}{2}}} \quad \ldots\ldots \quad (12)$$

where r_e = the emitter resistance.

It follows that from the aspect of both maximum power gain and stability there is little to choose between the common base and common emitter circuit configurations,

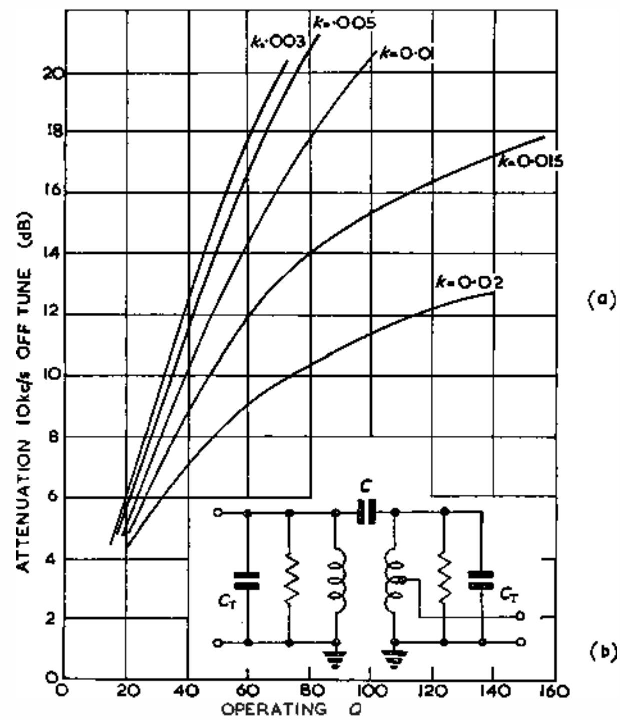

Fig. 10. Adjacent channel attenuation for double tuned coupling circuit

assuming neutralization is adopted in each case, and hence other receiver requirements such as selectivity or the transformer losses tolerable will usually decide the preferred arrangement. As the output resistance of the common base amplifier is always greater than that of the corresponding common emitter circuit, and both decrease rapidly as the 'α' cut-off frequency is approached, it is usual to find that the common base arrangement is preferred when the working frequency is near to $f\alpha$, say $f \geqslant \frac{1}{2}f\alpha$, but the common emitter circuit for frequencies below this range.

For currently available alloy transistors having 'α' cut-off frequencies of the order of 5Mc/s, practical stage gains of about 28 to 32dB in common emitter circuits are fairly typical.

Another important aspect of the i.f. amplifier design is the choice of coupling arrangement. In general this takes the form of a tuned circuit in the collector with a step-down winding to the following base or emitter input circuit. For minimum insertion loss the transformation ratio is chosen to match the output resistance of one stage to the input resistance of the next. The unloaded Q value should then be made high enough to provide the correct working Q for the overall selectivity required. This arrangement often suffices for lower quality receivers of the portable type, but some elaboration is usually required for, say, car receivers, where adequate bandwidth and better selectivity are necessary. In this case at least one double tuned coupling circuit will be required.

If double tuned circuits are employed the increased selectivity is usually obtained at the expense of increased circuit losses. Curves relating the relative attenuation 10kc/s off tune with operating Q are plotted in Fig. 10 for various values of coupling coefficient. For the circuit shown the coefficient of coupling, k, is given by:

$$k = C/(C + C_T) \quad \ldots\ldots\ldots\ldots\ldots (13)$$

The emitter of the following stage would normally be fed from a suitable tap on the secondary coil.

For the above type of circuit (or its equivalent using mutual inductance coupling) it has been shown[9] that the

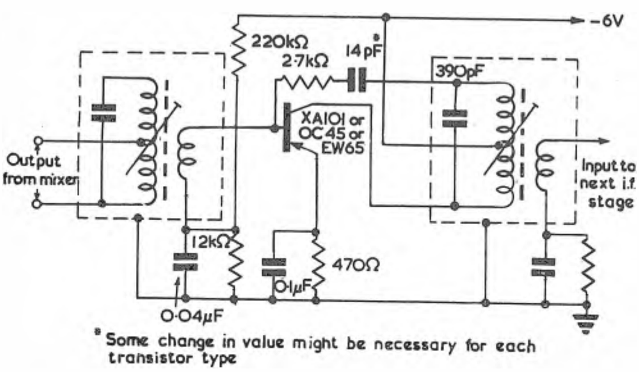

Fig. 11. Common emitter i.f. stage with bias stabilization

insertion loss G_I is given by:

$$G_I = 20 \log \frac{1 + k^2Q^2}{2kQ(1 - (Q/Q_o))} \quad \ldots\ldots\ldots (14)$$

Assuming $Q \simeq (Q_o/2)$, $Q_o \simeq 120$, and that an adjacent channel attenuation of 12dB is required, the insertion loss is found to be of the order of 7dB. These figures are probably fairly typical of those required in practice.

It is assumed that the operating Q's of both tuned circuits are equal.

Other types of interstage coupling circuit are possible, and these have been discussed in some detail in a paper by C. C. Cheng[10].

A good i.f. amplifier design should be reasonably independent of variations in the d.c. operating conditions. Such variations are mainly due to (a) the emitter bias current changing with variation of the current gain factor α_{cb} between different transistors, and (b) change of h.t. voltage which will affect both the emitter bias current and d.c. collector voltage.

To minimize the former some form of bias stabilization is necessary; a suitable circuit arrangement can be similar to that shown in Fig. 11. In the latter case the transistor's collector capacitance is dependent on the value of the collector voltage. This variation is governed in the case of alloy transistors by the relationship

$$C_c = K/\sqrt{V_c} \quad \ldots\ldots\ldots\ldots\ldots\ldots (15)$$

where K = a constant. Hence it is usual to use a large value of fixed tuning capacitance to swamp variations in C_c. In addition reduction of h.t. voltage will tend to reduce the emitter current, and hence reduce the loading on the preceding tuned circuit. Two i.f. response curves for a particular receiver are shown in Fig. 12. These curves were plotted for a 2:1 change in h.t. voltage, using a total tuning capacitance of 1 000pF, and refer to the total tuning characteristic of two cascaded i.f. stages preceded by a mixer. It will be seen that small changes have occurred both in the tuning and the selectivity characteristic of the amplifier, but they are not unduly serious.

A final consideration in the i.f. circuits is that of choice of frequency. As transistors having 'α' cut-off frequencies

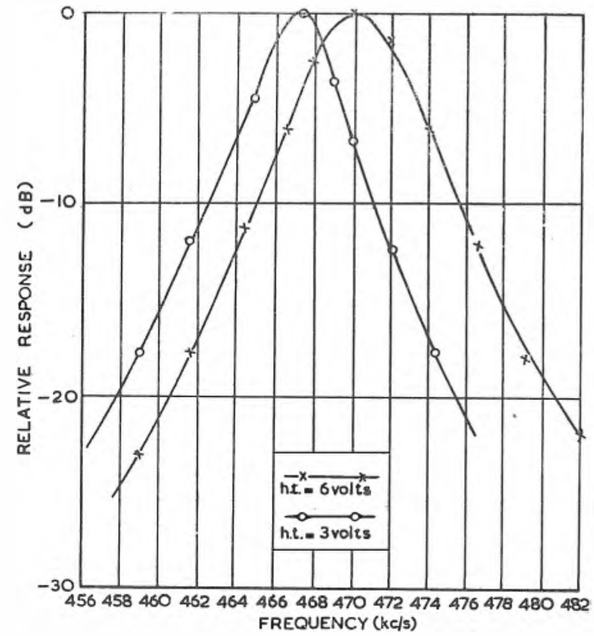

Fig. 12. I.F. response curves (for receiver 2 of Table I (Part 2))

greater than 5Mc/s are now available, there seems to be no strong reason why the standard i.f. of 470kc/s should not be used. This frequency is considered to be the best one on the basis of the existing channel allocations within the band. In this connexion it should be remembered that in the case of transistor receivers oscillator radiation may be more serious than in the case of an equivalent valve receiver, at least in the battery portable class. Image rejection troubles may also be more apparent.

Measurements of maximum oscillator radiation in the case of three transistor portable receivers have indicated a large spread, values of electrical field strength received on a 4ft rod aerial placed at a distance of 6ft from the receiver ranging from about 30μV to 300μV for the three models. A number of factors probably contribute to this rather large spread, not least the design of the ferrite aerial coil at the front end.

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(To be continued)

An Improved F.M. Discriminator

By V. B. Hulme*

An improved f.m. discriminator of the constant-area type has been developed. The circuit has been used in several applications where a frequency demodulator, having a linear characteristic over a wide deviation band, has been required. It has been proved to possess the advantages of simplicity, reliability and superior noise performance, when compared with other examples of this type of discriminator.

AN improved f.m. discriminator circuit which contains only five double valves and which provides a linear relation between input frequency and output voltage with non-linear distortion of the order of 0·1 per cent, over the range 100kc/s to 1Mc/s is described. A cathode-coupled multivibrator is employed as a 'constant-area' current pulse generator. High d.c. stability is achieved by means of negative current feedback, and operation at high frequencies is made possible by shaping current rather than voltage pulses, and by the use of diodes for the rapid restoration of initial conditions at the beginning of each pulse cycle. Amplitude modulation of the input signal has negligible effect upon the output.

The Demodulator Circuit

F.M. demodulator circuits fall into one of two classes, viz. those like the Foster-Seeley discriminator which rely on the phase-shift associated with a resonant circuit, and those which comprise pulse circuits of the integrating or counter type. Demodulators of the latter type are generally complicated and have many valves, but they give superior performance with wide modulation bandwidths. This type may be further sub-divided, distinguishing between circuits whose operating cycle is complete within one period of the input frequency, and others which average the frequency of several cycles of the input over a longer period. The circuit described is of the former sub-type.

The signal voltage (Fig. 1(a)) is amplified, limited and differentiated in such a way as to produce positive trigger pulses (Fig. 1(b)) at times corresponding to one of the zero intercepts of a cycle in the signal waveform. These pulses trigger a circuit which produces a current pulse (Fig. 1(e)) of constant amplitude and duration for every cycle of the input frequency; the voltage developed across a resistive load by this current after it has traversed a low-pass filter is the output of the discriminator.

Constant-Area Pulse Generator

The constant-area pulse generator is a cathode-coupled flip-flop in which V_2 is normally conducting and V_1 cut off (Fig. 2). The behaviour of this circuit is well known, and it is easy to see that, for each cycle of the signal, a constant quantity of electricity IT_D ceases to flow in the anode circuit of V_2; thus a voltage across the load resistance proportional to signal frequency and equal to $IR_1(T_D/T - 1)$ is developed, where $T = 1/f$, and R_1 represents the parallel combination of the two terminating resistors of the low-pass filter which is employed to remove the pulse frequency components in the output voltage. Since V_2 is virtually a constant current generator, its output impedance is high and may be neglected when matching the filter impedance.

Advantages of the Circuit

The basic circuit described is capable of giving an output voltage linearly related to signal frequency from a very low frequency to a maximum frequency, f_{max}, whose period, T_{min}, is equal to the pulse length T_D added to the recovery time T_R. The sensitivity in volts per cycle per second will be $IR_1/[f_{max}(1 + T_R/T_D)]$. For efficient operation at high frequencies, the recovery time T_R must be as short as possible. Without the diode V_3, the recharge of C_2 would be exponential and would require an indefinite recovery period; the frequency limitation would be severe and the slope of the demodulation characteristic would diminish continuously as the operating frequency increased. With suitable values of the circuit components of Fig. 2, a recovery time of 0·2μsec is possible and this permits linear operation up to 1Mc/s. It should be noted that the voltage waveform at V_2 grid, as shown in Fig. 1(f), when the pulse length is made small so as to obtain a high maximum frequency, is rounded because the flip-flop transition period is then comparable to the pulse length.

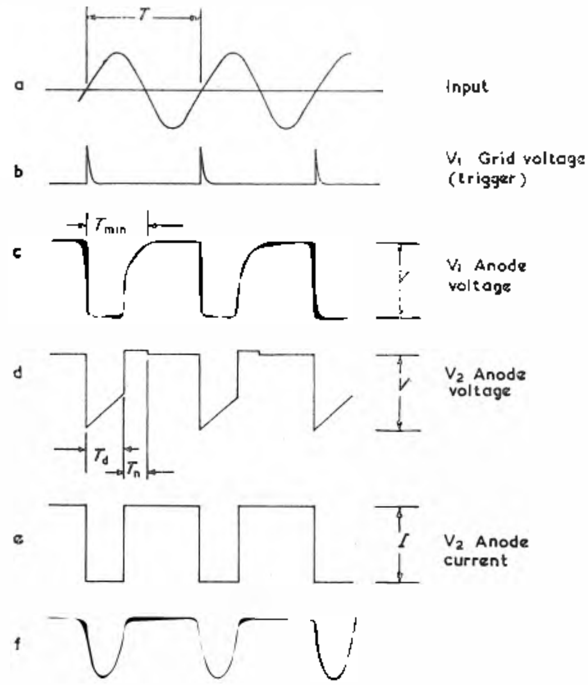

Fig. 1. Waveforms

An important feature of the circuit is the negative current feedback which occurs in V_1 or V_2 cathode circuit when only one of the pair is conducting. The feedback factor g_mR_3 has been made as large as conveniently possible by allowing a large proportion of the h.t. voltage to appear across the cathode load. Thus, the voltage amplitude V is stabilized and helps to provide a stable pulse length T_D and pulse amplitude I; the effects of changes in valve conductance are reduced by the factor $1 + g_mR_3$, which is approximately twenty times. The effect of varying the amplitude of the trigger pulse is also reduced by a similar factor.

* E.M.I. Electronics Ltd.

Derivation of the Trigger Pulse

As already noted, the trigger pulses should be related in time to one of the zero intercepts of the signal, so that amplitude modulation does not introduce spurious phase modulation of the trigger pulses which will be interpreted by the demodulator as frequency modulation, and thus give a spurious output (see Appendix). It should be noted that there is little restriction on the waveform of the signal.

The time relation should be such that the delay between the time of zero intercept and the firing of the constant pulse generator will be a constant or zero, independent of signal amplitude. To achieve this condition, it is necessary for the gain of the limiting stages to be sufficient to provide a transition of adequate slope even when the input signal is quite small. Any delay in reaching the voltage needed to trigger the flip-flop due to stray capacitance should be negligible or stabilized. In the present circuit, the delay has been stabilized at the point where the trigger pulse is shaped, and minimized elsewhere by attention to the high frequency response of the amplifier.

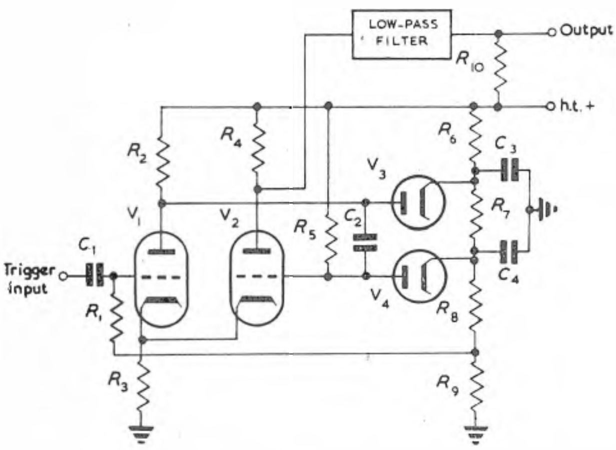

Fig. 2. Constant current generator (flip-flop)

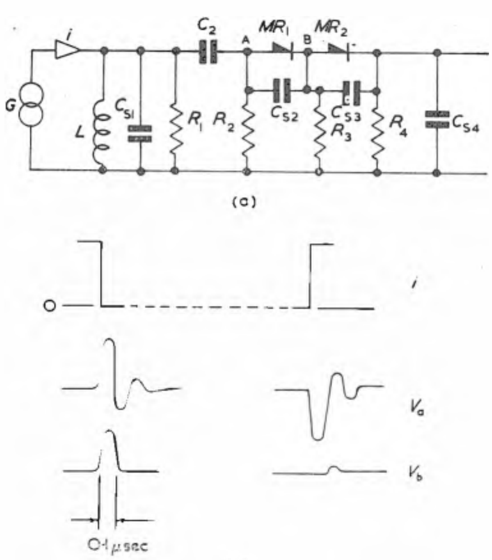

Fig. 3. (a) Pulse shaping network (b) Waveforms

The equivalent pulse shaping circuit, which is shown in Fig. 3, consists of an *LC* ringing circuit followed by two crystal diodes which suppress the negative part of the waveform. During half of the signal period, a steady current from the current generator builds up a magnetic field associated with the inductor L_1. When the current is interrupted, the field collapses and an oscillatory current circulates through L_1 and R_1, and the stray capacitances produce a voltage with a damped oscillatory waveform.

The Limiting Amplifier

The current of square waveform required to drive the pulse shaping circuit is produced by a cathode coupled amplifier since this type of amplifier lends itself to limiting at potentials which are symmetrical with respect to zero and is substantially independent of the input amplitude. With other circuits, it is generally difficult to avoid unidirectional currents which introduce spurious d.c. components and shift the 'slicing level' from the zero line. In the circuit of V_2 (Fig. 4), the mean cathode voltage is arranged to allow reasonably efficient operation of the circuit as a long-tailed pair. A consideration which fixes the lower limit to the value of R_k is the necessity of avoiding grid current in either valve with large input voltages.

The voltage gain of the amplifier/limiter stage V_1, without the regenerative coupling R_2C_3 is approximately $g_mR_4/2$, if $R_a = R_4 + R_3(1+\mu) \gg R_4$ and $R_3 \gg 1/g_m$ and, since the maximum value of R_4 is limited by stray capacitance, a gain of between five and ten would be realized. However, a useful increase in gain (two or three times) is obtained by the use of positive feedback from the first anode to the second grid as shown; the resultant gain is approximately $\dfrac{g_mR_4/2}{1-g_mR_2/2}$.

It should be noted that this circuit becomes self-oscillatory when $g_mR_2 > 2$ and that the condition for infinite gain and, hence for minimum input voltage swing required to switch V_2 on and off, is $g_mR_2 = 2$; this condition is independent of the value of the load R_4. The latter may therefore be replaced by the pulse shaping circuit of Fig. 3 in the amplifier/limiter stage V_2.

To suppress outputs resulting from amplitude modulation, the minimum signal to be accommodated must cause the current switch-over time to coincide with the signal zero intercept, within a tolerance dictated by amplitude modulation effects (see Appendix). This implies that adequate amplification should be provided before limiting takes place. If the permissible departure from zero level clipping is a fraction $\delta V/V$ of the carrier voltage level and the normal variation of the clipping level (a characteristic of the limiter circuits) is ΔE volts, the amplifier must provide an output of $V\Delta E/\Delta V$ volts without distortion, with a voltage gain of $V\Delta E/V_{e(min)}$ where $e_{(min)}$ is the amplitude of the smallest signal to be accommodated. These conditions must be satisfied over a wide range of signal levels, say 40dB, so that in practice it will not be possible to say which stage of a multi-stage amplifier the limiting process begins. Consequently every stage should have similar characteristics and satisfy the requirements both as an amplifier and as a limiter.

The Complete Discriminator Circuit

The output voltage at the anode of V_{3b} (Fig. 4) contains the unwanted carrier frequency and harmonics as well as the modulation frequencies, including a d.c. component. A low-pass filter is employed to suppress the carrier frequency components and to limit the modulation bandwidth. A choice of two bandwidths is provided by alternative filters and a switch so that the maximum signal-to-noise ratio may be realized with nominal cut-off frequencies of 10 and 40kc/s. The filter coils are given suitable *Q* values to avoid over-shoot on modulation transients

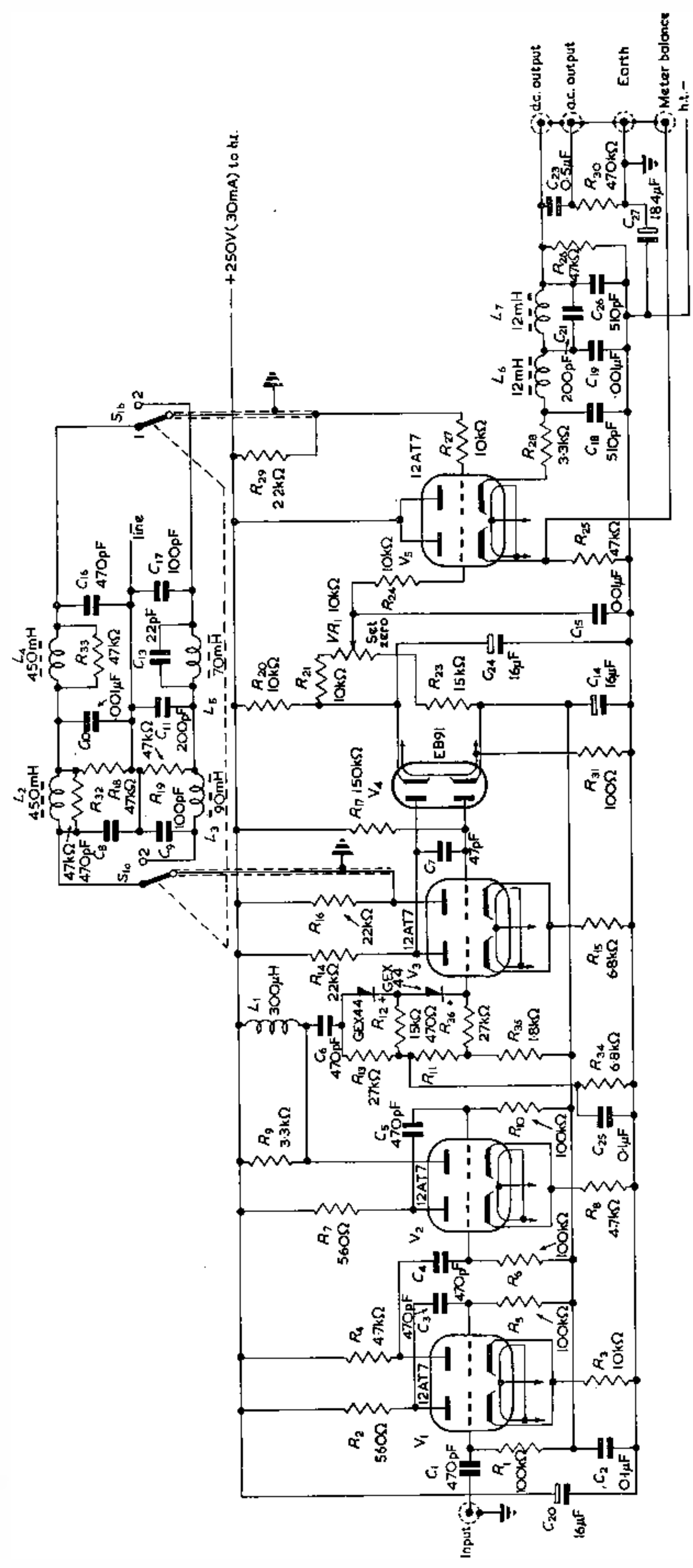

Fig. 4. Circuit diagram of discriminator

and the rise-time of response to a step function is approximately equal to a period of the cut-off frequency.

The output from the filter is directly coupled to a cathode-follower which is matched into a low impedance filter for further attenuation of carrier frequency components; this gives an overall attenuation, due to both filters, exceeding 100dB. This is necessary to ensure that carrier frequency components in the output are 60dB or more down relative to the modulation when the carrier deviation is say 10 per cent of the carrier frequency. The output end of the cathode-follower filter need not be matched to the load since its cut-off frequency is higher than that of the preceding filter, which protects it from high frequency components and transients which might otherwise cause overshoot and ringing.

The second cathode-follower has a variable d.c. bias applied to the grid via a potentiometer, and the low impedance output is used to balance out the unwanted d.c. component of the discriminator output. The potentiometer control is accessible to the user and is labelled 'set zero'. By earthing the balance output instead of the h.t. negative rail, the discriminator output may be adjusted to give d.c. zero at any carrier frequency in the working range.

The h.t. power supply, which is not shown in the diagram, provides a stabilized supply of 30mV at 250V. The h.t. ripple voltage must be very small (less than 1mV peak-to-peak) so that it will not appear in the discriminator output.

The noise contributed by the discriminator itself is too small to be observed although at one stage in the development, the noise became observable if the thermionic diodes were replaced by crystals. The major part of the noise which could be attributed to the discriminators when operating under poor conditions is likely to be due to amplitude modulation of the carrier. However, the output due to a.m. is so small that it cannot be measured with available a.m. generators because of the masking effect of phase or frequency modulation in the a.m. generator output. Using an a.m. generator with a quartz-crystal master-oscillator and with 10 per cent amplitude modulation at 10kc/s, an output of 8mV peak-to-peak was obtained from the discriminator; but since it was not known how much of this was due to the presence of the modulating frequency or of phase modulation in the output of the generator, it is not possible to say how much is due to imperfections in the discriminator. A discriminator output of 8mV is 0·5 per cent of that due to 16kc/s f.m. deviation.

Acknowledgments

The Author wishes to thank the Directors of E.M.I. Electronics Ltd. for permission to publish this article.

APPENDIX

EFFECT OF AMPLITUDE MODULATION AND INTERFERING SIGNALS

In Fig. 5, two sine waves y' and y'' are drawn where

$$y' = E_{max} \sin 2\pi t/T \text{ and } y'' = (E_{min}/E_{max})y'$$

so that y' and y'' define the excursion of an a.m. carrier having a modulation index $M = (E_{max} - E_{min})/(E_{max} + E_{min})$.

The phasing of the discriminator constant-area pulse may be related to the intercept of a voltage level K with the input waveforms y' y'', at t_1, t_2 during, say, the region of positive slope in the cycle.

The discriminator will give an output, corresponding to phase modulation, of $2\pi(t_2 - t_1) = \Delta\theta$ peak-to-peak radians; i.e. a spurious output, due to amplitude modulation, which is $\Delta\theta/m$ times the normal output due to a frequency modulated signal having a deviation ratio $m = \Delta f/f_s$, where Δf is the frequency deviation and f_s the modulation frequency.

Now,

$$\Delta\theta = \sin^{-1} K/E_{min} - \sin^{-1} K/E_{max}$$

and if $K/E < 0{\cdot}1$

then $\Delta\theta \simeq \dfrac{(E_{max} - E_{min})}{(E_{max})\ (E_{min})} K$ with an error less than 0·2 per cent and, if $M \ll 1$,

$$\Delta\theta \simeq 2MK/E, \text{ since } E_{max} + E_{min} \simeq 2E$$

$$\text{and } (E_{max})(E_{min}) \simeq E^2$$

As an example, let $M = 0{\cdot}03$, $K/E = 0{\cdot}1$, $m = 2$, then the ratio of spurious output to the signal is $2MK/mE = 0{\cdot}3$ per cent.

The effect of an interfering waveform may be analysed by considering the vector diagram Fig. 6. The vector OA represents the desired carrier of frequency f_c and unit amplitude. AB represents a sinusoidal interfering signal of frequency f and relative amplitude r, and rotates with respect to OA at a difference frequency $(f_c - f)$, so that

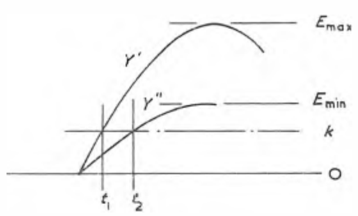

Fig. 5. Effect of amplitude modulation

the resultant OB has a phase modulation $\Delta\theta$. If $r \ll 1$, $\Delta\theta$ peak $= r$ and the undesired output is

$$(rA/\Delta f)(f_c - f) \cos 2\pi (f_c - f)\, t$$

where A is the peak output from a desired signal modulation of peak deviation Δf.

If the frequency of the interfering wave is very low e.g. outside the carrier band but in the signal modulation band, it is preferable to regard it in effect as a periodic variation of the level k, Fig. 5. The phase modulation for r small is $r\cos 2\pi ft$, and the output $rA/\Delta f\ f\cos 2\pi ft$.

Because the interfering wave and the level k does not vary appreciably during a half-cycle of the carrier frequency, since $f \ll f_c$, it is possible to reduce, if not eliminate, its effects (and also the effects of amplitude modulation

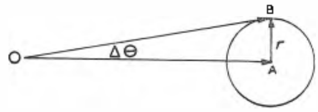

Fig. 6. Effect of interfering waveform

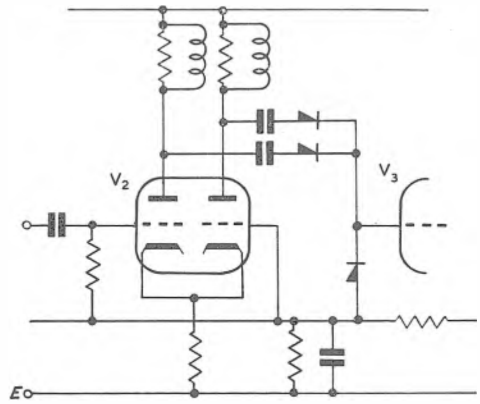

Fig. 7. Derivation of trigger pulse from each intercept

at low frequency) by triggering the constant area flip-flop on both intercepts, i.e. twice per cycle. The phase errors of the pulse pairs are then in opposition and in demodulation they cancel. The circuit V_2V_3 (Fig. 4) can be changed as in Fig. 7 to derive a trigger pulse from each intercept.

The arrangement has other advantages, e.g.:

(a) It avoids the risk of a discontinuity in the demodulator output/frequency characteristic at $f_c = 1/2T$, due to residual break-through of the suppressed negative trigger pulses.

(b) The effective carrier frequency is doubled at the carrier filter, making suppression easier.

An Electronic Telephone Exchange

One of the first electronically controlled automatic telephone exchanges to go into service as a practical working unit in Britain, began operating recently at Taplow Court, Taplow, (Bucks), formerly the home of the late Lord Desborough, and now the headquarters of British Telecommunications Research Ltd.

Designed to cater for 200 extensions, the exchange employs the magnetic drum storage technique giving unique facilities and will handle all internal communications within the organization.

Its development as a practical working unit is part of the research programme into electronic telephone switching, at present being carried out by B.T.R. Ltd, an establishment jointly sponsored by Automatic Telephone & Electric Co. Ltd, and British Insulated Callender's Cables Ltd.

All information received from telephones connected to the system is stored on the magnetic drum which decides the most suitable routes to establish a connexion, records the state of the trunking of the exchange. and also provides a register for each subscriber, a facility which would be economically impractical without electronic control techniques.

The exchange is designed to provide for calls to be booked to a busy subscriber. If the dialled number proves to be engaged it is only necessary to hang up again and when the called number becomes free, the exchange will remember the wanted call and automatically ring both calling and called subscribers. Such a facility is more appropriate to a private exchange than a public exchange, but gives an example of the extra facilities which can be provided quite cheaply by electronic control techniques.

The Design of the Ferranti Pegasus Computer

(Part 2)

By G. Emery*, M.A.

PART 1 of this discussion dealt with the considerations affecting the system design of a medium-sized general-purpose computer: Part 2 is concerned with the more detailed engineering aspects of the problem, and in particular with the implications of building the machine from standard plug-in packages.

Packaged construction leads to a number of economies, which become even more marked if the majority of the packages used are of a relatively small number of types. Economies arise

(a) In design because, the packages once having been developed, the machine can be designed almost entirely from a logical standpoint without the need for further laboratory work.

(b) In manufacture, particularly if quantity production is envisaged and the number of package types is small.

(c) In commissioning because, with pre-tested packages, the number of possible faults is greatly reduced.

(d) In maintenance, because for a relatively small number of package types a correspondingly small quantity of spare equipment need be held.

(e) In operation, as the time during which a fault makes the computer inoperative is reduced to the time taken to locate the fault, the faulty package then being replaced and removed elsewhere for repair.

Two disadvantages have been alleged against packaged construction; the first is that it leads to a number of redundant components; the second is that it increases the number of plug-and-socket connectors through which signals must pass, with a possible decrease in reliability. With the package types adopted, it is mainly the cheaper components —resistors and crystal diodes—that become redundant, their cost being far outweighed by economies in manufacture. The plugs and sockets have proved to be entirely reliable. The sockets are of the type used in standard octal valve bases, mounted in line; the plug pins are simply plated metal strips folded around a plastic moulding.

Package Types

Inevitably there must be several package types associated with special functions within the machine, of which only one or two will be needed in any installation. The majority of the machine, however, will be concerned with arithmetical and logical operations on digit-pulse trains, and can be made up from a relatively small number of package types. The basic operations required are the logical AND and OR operations and the NOT operation (inversion) and delays of one digit-time and one word-time. There is also a requirement for cathode-followers when the loading on a particular signal is heavy. The AND and OR operations can be conveniently performed by crystal gates; the other operations require some form of valve circuit. Thus the basic packaged elements can be envisaged as valve circuits fed from crystal gates or combinations of crystal gates. A convenient package size will accommodate up to three miniature-valve bases and a thirty-two-pin plug, allowing from two to six identical, or closely similar, elements per package.

Typical crystal AND and OR gates are shown in Fig. 1. It was decided to predesign all AND gates by building them on the packages and choosing a value of bleed resistor suitable to the succeeding valve circuit. OR gates are made up for the most part from 'mix' crystals provided as alternative outputs for the packaged elements; a choice

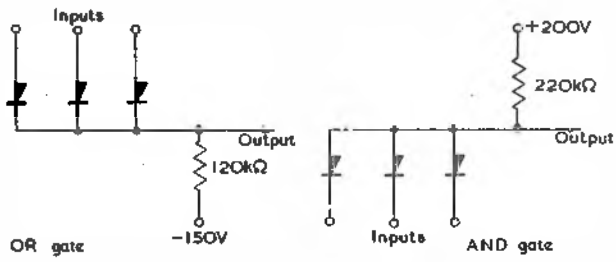

Fig. 1. Typical AND and OR gates

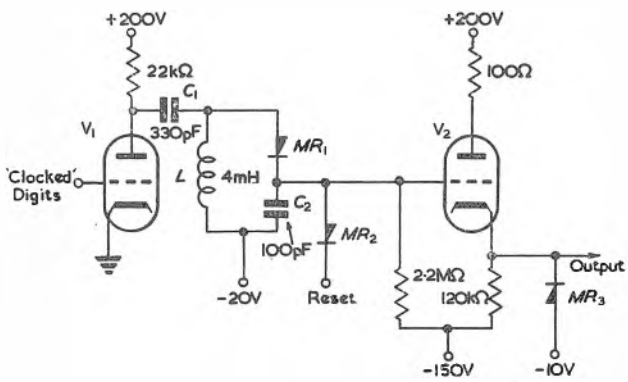

Fig. 2. Basic delay circuit

of bleed resistors, connected to the negative line, is provided on each package, the ends of the resistors being brought to pins on the package plugs. These resistors are also available as extra loads for the package output valves, which are always cathode-followers, to enable the logical designer to adjust each functional element to suit the external load imposed on it.

The packages are designed for a pulse frequency of ⅓Mc/s. The digit pulses are positive-going, and consecutive pulses are separated by gaps of about 0·6μsec. They are generated in the first instance in the digit-delay circuit by the use of two auxiliary waveforms, 'clock' and 'reset'. 'Clock' is a ⅓Mc/s square wave of unity mark-to-space ratio: 'reset' is a negative-going ⅓Mc/s pulse of width 0·6μsec. The basic digit-delay circuit is shown in Fig. 2. It accepts a 'clocked-up' digit pulse (i.e. a pulse that has undergone an AND operation with 'clock' to standardize its timing); this causes a current to be built up in the inductor, which is diverted to capacitor C_2 when the input level falls. The output is now positive until the end of the next digit-time when C_2 is discharged by 'reset and the output falls again. Owing to the degenerative effect

* *Ferranti Ltd.*

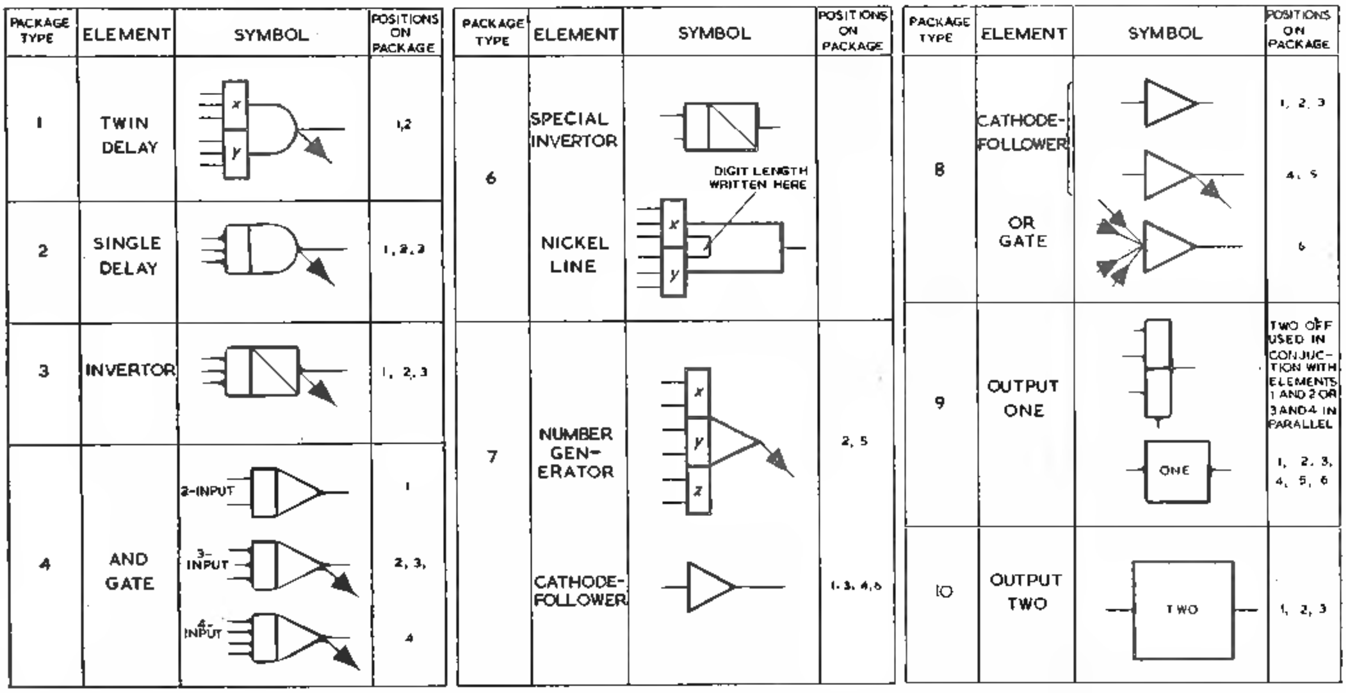

Fig. 3. Packaged logical elements

of cathode-followers, which must be inserted between gating operations, it is necessary to re-form digit pulses in this way at intervals in their travel. Because pulses are 'clocked-up' before entering delays, only the part of a waveform 'overlapped' by 'clock' has any real significance; consequently inter-digit spikes produced by logical operations between digit-pulse trains with slightly different timing can have no effect on the operation of the machine.

Digit-delay packages are of two types: in one the delay is fed from an AND gate, in the other from the mixed outputs of twin AND gates. This so-called 'twin' delay has the facility that it can be used as a staticizer, a function for which flip-flops have more generally been used. This is done by feeding the output of the element back to one of the input gates; a pulse injected through the other gate will then circulate until the feedback gate is closed.

The other packaged logical elements are invertors—a simple triode inverting amplifier direct-coupled to a cathode-follower and preceded by an AND gate—AND gates feeding cathode-followers, OR gates feeding cathode-followers, and cathode-followers alone. There is also a dynamicizer element consisting of three two-input AND gates with their mixed outputs feeding a cathode-follower; this is intended for generating serial numbers from hand-key settings or tape-reader outputs. Finally there is a delay-line package, which is used primarily for immediate-access storage. The line material is nickel, which has a dispersion characteristic and thermal coefficient of delay sufficiently low for the lengths used, and in which a sonic pulse can conveniently be induced by magnetostriction. The pulses received at the remote end of the line are amplified and then reshaped in a circuit identical with those on the digit-delay packages, to give a standard digit-pulse output over a temperature range of some 20°C. The amplifier at the transmitting end of the line is fed from twin AND gates, one gate being used for filling the line, and the other for recirculation. The line and its associated coils are mounted on a panel that can be easily removed from the package and replaced if necessary.

Fig. 3 shows how the packages are made up from the basic elements. It will be seen that most elements have the alternative of direct and mix-crystal outputs. Types 9 and 10 are intended for controlling output equipment; the type 9 package also carries twin AND gates, which are useful when the element is to be used for setting and resetting relays. Some examples of the use of the packaged elements are shown in Fig. 4. The parity counter is essentially a staticizer that is cut on and off alternately by the pulses in the input signal; it is set at the beginning of the cycle, and its output at the end of the cycle is positive for an even-parity signal. Parity counts are used as checks at several places in the computer. The adder-subtractor shown is the most convenient with the package types adopted; a delay is used at the output point to give a standardized 'sum' signal.

The packages are built on plates of a plastic laminate, 12in long and 6in wide, with a cut-out arranged to form a carrying handle. Within the cut-out is an aluminium platform, which is mounted at right-angles to the main plate and will accommodate up to three miniature valve bases and up to eight monitor-probe sockets. When the packages are mounted in the cabinet these platforms form a division that separates the cooling-air stream over the valves from that over the rest of the package.

The Drum Store

The main backing store in Pegasus is a drum of diameter 10in rotating at 3 720rev/min. To record digits at the standard 3μsec rate would imply a cell length of about ·006in, which might not have given the desired reliability; consequently digits are recorded at 6μsec, a pair of tracks being used at a time to record alternate digits. The method of recording is the so-called phase-modulation system, in which a binary 'one' is represented by magnetic saturation for 3μsec in the negative direction followed by magnetic saturation for 3μsec in the positive direction, a binary 'nought' being represented by the inverse of this. The logical circuit shown in Fig. 5 generates suitable phase-modulated signals and provides double-ended outputs for feeding the push-pull write amplifiers. Gating waveforms

T_3 and T_4 select odd and even digits respectively from the original signal. The read signal derived from the drum is strobed to eliminate timing variations (the 'reset' waveform is used for this) and gated again with T_3 and T_4 to reconstitute the original information.

The drum is provided with one hundred write/read heads, fitted in ten stacks of ten. Eighty heads are used for storage purposes, giving a maximum storage capacity of forty track pairs, of which eight pairs are used for a permanent record that can be read from but not, in normal circumstances, overwritten. Four heads are used for two address-track pairs, and three heads are used for clock tracks, which provide synchronism between the drum and the rest of the computer. The address and clock tracks are duplicated to allow the permanent record to be written without the necessity for reading and writing with two heads on the same stack. The third clock track is necessary to enable the other two to be written in exact synchronism. The remainder of the heads are treated as spares.

The drum clock track provides a sinusoidal signal which is amplified and shaped to give the 'clock' and 'reset' signals that control the digit-pulse frequency throughout the computer. It is additionally necessary to keep the drum-clock frequency constant within narrow limits to ensure that signals in the fixed-delay nickel lines are not spoiled. The drum-clock frequency is therefore compared with the output of a crystal oscillator, the error being used to control the field current of the generator supplying the drum motor. For testing without a drum, 'clock' and 'reset' can be generated direct from the output of the oscillator.

Each track pair carries 5 376 digits (128 words or 16 blocks); thus the total storage capacity is 5 120 words. A word or block is selected from a track by address-track coincidence. The tracks themselves are switched electronically on the special packages designed for reading and writing. For this purpose the heads are assumed to be arranged as matrices of eight rows and five columns; there are two identical matrices, one for the odd digits and one for the even digits. A column generally corresponds to eight heads in the same stack; these are connected to eight transformers on a special switch package, which are switched to the read amplifier by suitable low-level selection signals. The switch package also carries a buffer amplifier, which presents to the read amplifier a signal substantially the same as would be obtained direct from the head; thus the same read amplifier circuit can be used for the information tracks as for the address tracks, where there is no switching. Column selection for writing is carried out at a low level by gating operations at the inputs to the write amplifiers. (There are only four write amplifiers for odd digits and four for even digits, as the permanent record occupies one complete column.) Each head transformer carries a third winding connected so that the eight transformers on one switch package provide a choice of anode loads for one write amplifier; row selection within a column for writing is then achieved by applying the h.t. for the write amplifier to a centre-tap on the winding of the appropriate transformer.

To ensure that words on the drum have a fixed timing relative to words in the immediate-access store, the commutator, which provides p-pulses at word frequency for use

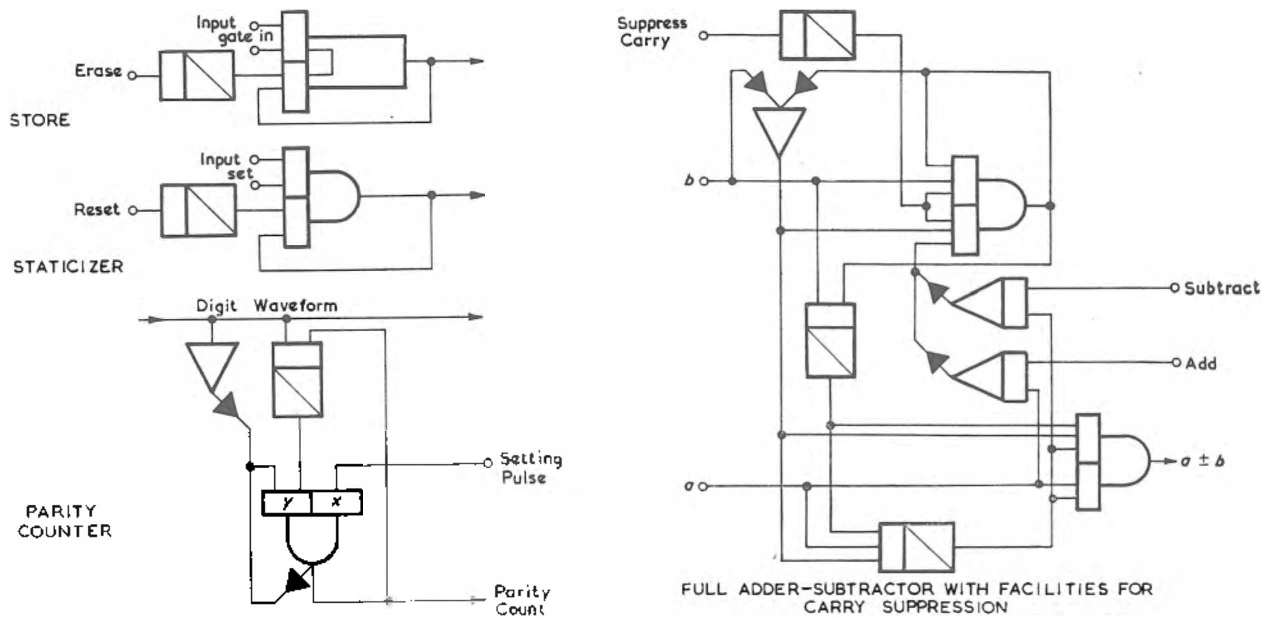

Fig. 4. Common logical circuits

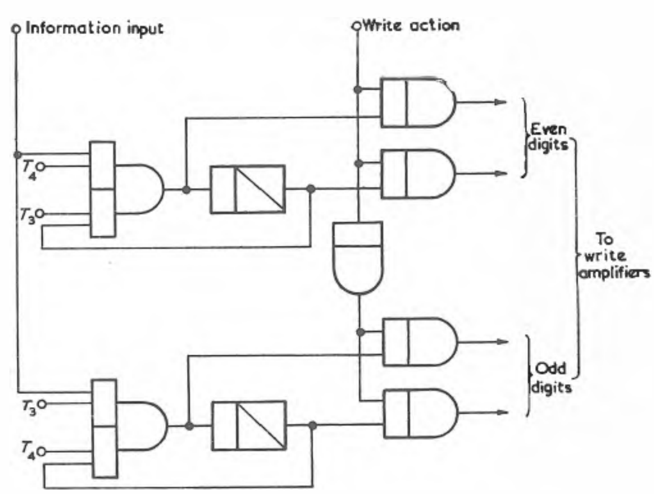

Fig. 5. Phase modulator

throughout the computer, is locked to a signal on the address track. The commutator is shown in Fig. 6. Essentially it consists of three loops cycling with periods of 7, 3 and 2 digit-times respectively, giving a combined cycle time of 42 digit-times, or one word-time. The arrangement is such, however, that the overall cycle time is 42 digit-times only if the signals from the pair of address tracks, H_8 and H_9, are both positive at the beginning of a new cycle, when the outputs of all delays in the loop are negative. As long as this condition is not fulfilled, the cycle time of the commutator will be 43 digit-times. The address-track signal is such that H_8 and H_9 are positive together in one particular place relative to every word on the drum, so that the commutator will eventually become locked after a settling-down period of a few drum revolutions. The effect of the address-track signal is over-ridden when the computer is switched to operate on crystal-oscillator clock.

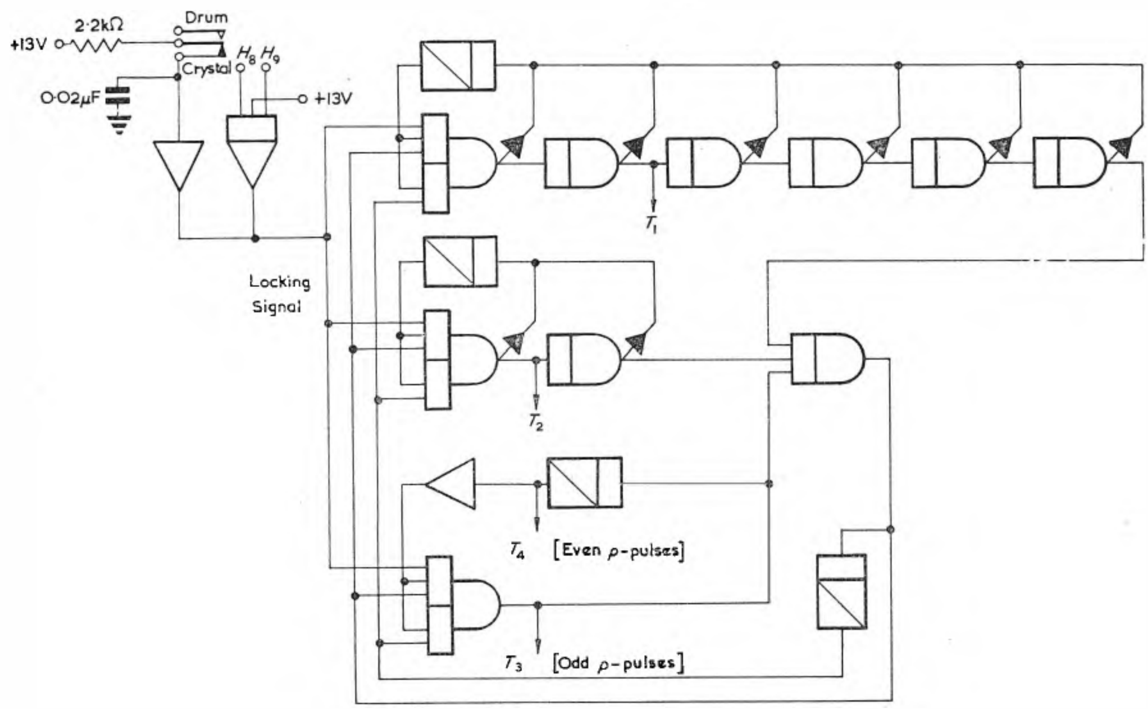

Fig. 6. Commutator

Arithmetical Circuits

As explained in Part 1 of this article, Pegasus is provided with multiple accumulators. To economize on equipment, it was decided not to associate a separate arithmetical circuit with each accumulator, but to carry out all simple arithmetical operations in a central 'Mill'. The delay necessarily involved in reshaping pulses during their passage through the Mill is padded out to one word-time to preserve standard timing in store; consequently orders must take at least two word-times to be obeyed. The separate Mill does give some saving in computing time, however, by enabling the source of either operand, i.e., any register in the immediate-access store, to be used for replacement after a simple arithmetical operation. The Mill is also used as a shifting register, and contains facilities for testing 'jump' conditions and for carrying out the arithmetic associated with the 'unit count' and 'unit modify' jump orders. Operands used in shift and jump orders are always taken from an accumulator.

Two of the accumulators are used as the basis of the multiplier-divider, and will contain the double-length product after multiplication or the quotient and remainder after division. As the order structure allows only two addresses to be specified, the less significant half of a double-length dividend must be placed in the multiplier-divider by a previous order. The two accumulators of the multiplier-divider cycle independently, and digits are shunted from one into the other during the course of an operation. During multiplication, the product is formed initially in the more significant of the two accumulators by means of three adders, three digits of the multiplier being sampled each word-time. It is possible to form the product three digits at a time because only 39 of the 42 digits in a word are arithmetically significant, leaving three gap digits into which the growing product can expand. Facilities are available for forming a double-length product or a rounded single-length product, or for accumulating double-length products.

Division is by a standard non-restorative process, the dividend being gradually reduced in numerical size by adding or subtracting the divisor (according to the sign of the residue) in successively lower significance, the quotient being built up in the other accumulator by corresponding half subtractions or additions. As the least significant digit of the quotient so far formed is a 'one' at every stage in this process, the operation must be carried one step beyond the end of the dividend to obtain the correct values of quotient and residue. For unrounded division (residue having the same sign as the divisor) the extra quotient digit is removed by a single half subtraction, a corresponding operation being carried out on the residue; for rounded division (minimal residue) the sign of the residue is sensed one step further still to determine the last operation. In either type of division the last two operations are carried out with the same significance, leaving a quotient and residue of standard length.

An interesting feature of the process is the method used to ensure that the final unrounded quotient and residue are correct when the residue has been reduced to zero at some intermediate stage. This condition is sensed; but the process is not terminated. Instead, an addition is forced on the quotient (and a subtraction on the residue) regardless of the sign of the divisor; the net effect of subsequent operations is then to restore the original values of quotient and residue while shifting them into standard timing. This special procedure is necessary, of course, only for unrounded division, though, for simplicity, it is allowed to operate for rounded division too.

The multiplier-divider circuit is used also for double-length shifting (including normalization for floating-point working, which is a double-length operation) and for justification. It comprises 54 packages, which is less than one eighth of the total package content of the machine.

Control

Following normal practice, the order pairs are routed in sequence (in the absence of 'jumps') into an order register, from which they are decoded. The first order is modified on entry; the second order when the operation prescribed by the first order is nearly completed, the first order then being deleted to allow the second order to be extended by modification. The adder used for modification is also used to step on the order address, so that order pairs are selected in the sequence in which they are stored, and for counting shifts, which are performed at the rate of one place per word-time. Orders are decoded by setting their digits up separately on staticizers of the type already referred to. The staticized outputs and their inverses are then combined to provide suitable gating waveforms. It should be noted that the logical arrangement of the function code, while providing a useful mnemonic for the programmer, sometimes enables the outputs of the function-decoding staticizers to be used directly as action waveforms in cases where several different functions require parts of the machine to be associated in the same way. Two sets of address decoding are used, one for the N-digits, which may specify any immediate-access address, the other for the X-digits, which specify one of the accumulators. The N-decoding is always used for replacement, leaving the X-decoding circuit free for setting up the modifier address for the next order. (It will be remembered that modifiers in current use are held in accumulators.)

Control can exist in one of three states: the C-state, during which the new order pair is entering the order register; the A-state, during which the first order is being obeyed; and the B-state, during which the second order is being obeyed. The C-state persists for one word-time, and the other two states for at least two word-times each. A beat counter provides signals to define the three states and to control the operations that are peculiar to each one. When computation is stopped, whether intentionally or due to a component failure or a programming error, the operation of the beat counter is suspended, so as to keep the order to be obeyed next circulating in the order register but unable to exercise control over the machine. The stop order and the optional stop (stop-go digit) operate in this way, also parity failure or the decoding of a function to which no significance has been assigned. Overflow does not of itself cause a stop; but the machine is stopped if there is a subsequent attempt to transfer information to the drum. Overflow can occur in the Mill or multiplier-divider after an addition, subtraction or left shift, or after a division operation in which the dividend numerically exceeds the divisor. Overflow can be cleared by either of the two jump orders that are conditional on the overflow setting, or by a 'justify' order.

Input and Output

Input and output are by five-hole paper tape, an interpreter unit being provided to give a print-out of the output tape. The input and output units are treated in orders as an address in the computing store, the two being distinguished by the function in the order according as it prescribes the address as a source or destination. Two addresses are however allotted to input and output. One of these implies a straightforward conversion from tape holes to digit pulses, and is generally used for reading in warning characters for routine interpretation; the character is read, for convenience, into the modifier position of an accumulator. When the other address is specified, the machine carries out an automatic conversion, replacing the most significant digit with a digit depending on the parity of the character. As all the odd-parity characters on tape are numerical or other characters used in programmes, a single-digit error in the reading mechanism, followed by the conversion, will cause to appear in the machine a character that is completely out of context. The input routine is arranged to test the significance of what is read in, and to enter a loop stop (a jump order transferring control to itself) if an error of this kind has occurred.

A great deal of time can be saved if alternative tape readers are available, one for reading in a master programme and the other for reading subroutines. It is a convenience, too, to have alternative tape punches. An external conditioning order allows a choice of up to nine pieces of tape equipment to be switched by relays operating under programme control.

The optional facility is provided in Pegasus for punching out addresses in drum block-transfer orders. A warning character is punched, followed by the ten digits of the block address in the form of two tape characters. The facility is particularly useful during programme development, and is achieved at the expense of very little extra equipment.

The Monitor

Monitoring facilities are required not only by the maintenance engineer but also by the programmer, especially during the development of a programme. Accordingly two c.r.t. displays are provided. One of these is referred to as the Programmer's Display, and will show either the current order pair or its address. The form of presentation is one in which binary 'ones' are shown as vertical lines, and binary 'noughts' as dots. The order pair is shown on a twin-line trace. The other display is referred to as the Engineer's Display, although many of the facilities associated with it are used by programmers too. All store lines are switchable to this display, as are several other crucial places in the machine, but it is possible to monitor the output of every packaged element in the machine on this display, as all packages carry monitor sockets, and probes connected to the monitor switch are provided in each computing bay. The left-hand display will give a presentation of the 'lines-and-dots' type used on the other display, or it will show 'clocked' digits or the raw digit-pulse waveform, the 'clocked' presentation enabling the engineer to determine whether any 'spikes' produced by the logic are occurring in significant positions relative to clock. A range of time-bases is provided, from the 'half word' length as used for the right-hand tube up to 30msec.

The time-bases are triggered as determined by the programmer in coincidence with a particular order address or drum address set up on hand-keys, or on a particular operational beat. Provision is made for triggering repetitively (which is generally required when an order is being run from the hand-keys for test purposes) or for triggering from an internal waveform derived from a probe in one of the computing bays.

Physical Layout

The computer comprises a three-bay computing cabinet with control desk attached, a two-bay power supply cabinet, and a motor-alternator set. The first bay of the computing cabinet, which abuts against the control desk, contains the drum and monitor unit and four rows of packages; the other two bays contain ten rows of packages each. The packages are mounted vertically between cast aluminium-alloy racks in rows of up to twenty. The sockets that mate with the package plugs are mounted in panels which divide the bays laterally into two parts, interconnexions between packages in the same bay being made by wiring between the socket pins in the back portion of the cabinet. Special plugs and sockets identical with those used for the packages are used for bay-to-bay interconnexions.

The lower portion of the cabinet, between the packages and the floor, contains heater transformers and cooling fans; the upper portion, above the packages, contains the inter-bay power wiring and protective circuitry. When refrigeration is fitted, which is necessary for the comfort of the operator if the computer is installed in a confined space, cooling batteries are mounted over the air intakes under a false floor at the front of the cabinet.

Acknowledgments

The author acknowledges his debt to all those who were responsible for the design of Pegasus. He would also like to thank Ferranti Limited for permission to publish this article.

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A Stabilized D.C. Power Supply Using Transistors

By T. H. Brown* and W. L. Stephenson,* B.Sc.

This article discusses the design of a mains operated d.c. stabilized power supply using transistors and incorporating negative feedback to provide a low output resistance. The output voltage is variable over the range 0 to 30V at currents of up to 1A, and the change in output voltage over this current range is less than 40mV.

THE increasing use of transistors at high current levels has led to a demand for a mains-operated stabilized power supply providing low voltage at currents of the order of 1A. Some aspects of the use of transistors as series stabilizers for this purpose have already been described[1,2] as well as feedback control systems using both a.c.[3] and d.c.[4] techniques.

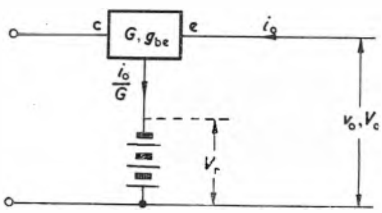


Fig. 1. Simple form of stabilizer

This article describes the theory leading to the design of a mains-operated stabilized power supply providing an output voltage which is continuously variable over the range 0 to 30V at a current of up to 1A. Attention is paid to the stability of output voltage with temperature and the provision of a low output impedance.

The simplest form of voltage stabilizer is the emitter-follower with the base connected to the reference battery as in Fig. 1.

The transistor is drawn here in block form with a current gain G and a base-emitter transconductance g_{be}. It may then represent either a single transistor or a number of transistors cascaded to form a compound emitter-follower.

* *Mullard Research Laboratories*

OUTPUT RESISTANCE

The differential output resistance of such a system may be calculated by considering that a small low frequency current is flowing into the system causing a voltage v_o to appear across the output. Then if r is the internal resistance of the reference,

$$v_o = -v_{be} + (i_o r/G)$$

and $i_o = -v_{be}g_{be}$

$$\therefore \; r_o = (v_o/i_o) = (1/g_{be}) + (r/G) \; \ldots\ldots\ldots (1)$$

When a single transistor is used,

$$(1/g_{be}) \approx r_e + \frac{r_b}{1+\alpha'}, \; G = 1 + \alpha'$$

With typical values at 100mA of $r_e = 0{\cdot}25\Omega$, $r_b = 25\Omega$, $1 + \alpha' = 50$, $r = 2\Omega$, $r_o = 0{\cdot}79\Omega$ and $G = 50$ and a comparatively large current (2mA) is drawn from the reference battery.

Using two transistors to form a compound emitter-follower as in Fig. 2, $1/g_{be}$ is now given by

$$(1/g_{be}) = r_{e1} + \frac{r_{b1} + r_{e2} + (r_{b2}/(1+\alpha_2'))}{1+\alpha_1'}$$

$$= r_{e1} + \frac{r_{b1}}{1+\alpha_1'} + \frac{r_{e2}}{1+\alpha_1'} + \frac{r_{b2}}{(1+\alpha_1')(1+\alpha_2')}$$

and

$$G = (1+\alpha_1')(1+\alpha_2')$$

At currents of more than a few milliamperes the d.c. current in X_1 is approximately $(1+\alpha_1')$ times that in X_2

$$r_{e1} \simeq \frac{r_{e2}}{1+\alpha_1'}$$

so that

$$(1/g_{bo}) \simeq 2r_{e1} + \frac{r_{b1}}{1 + \alpha_1'} \quad \dots\dots\dots\dots \quad (2)$$

giving $r_o = 1\Omega$ and $G = 2\,500$ at 100mA output current.

OUTPUT VOLTAGE

The d.c. output voltage in the general case is given by

$$V_o = V_r - V_{be} \quad \dots\dots\dots\dots\dots\dots \quad (3)$$

and for a single transistor V_{be} is generally less than 0·5V.

The variation of V_{be} with current has already been considered in terms of output resistance and due to the high

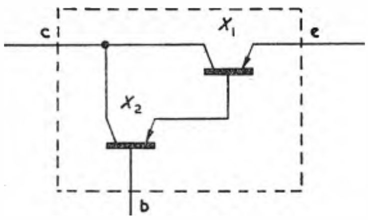


Fig. 2. Use of two transistors to form a compound emitter-follower

collector resistance, its variation with collector voltage is negligible, but at a fixed current V_{be} varies with temperature by 2·5mV/°C. Hence even with a constant load current the output voltage can vary with the transistor working temperature.

For a compound circuit using more than one transistor, this temperature variation increases in proportion to the number of transistors used.

The Use of D.C. Feedback

The relatively high output resistance of the emitter-follower circuits leads to a study of d.c. feedback to effect some reduction.

In general terms the system shown in Fig. 3 consists of an emitter-follower as before, but with d.c. feedback applied through a phase-reversing amplifier.

From general considerations, the presence of this feedback loop can be expected to reduce the output resistance of the compound emitter-follower by a factor of 1 + A where A is the voltage gain of the feedback amplifier. The effect on the output of variations in the unstabilized supply

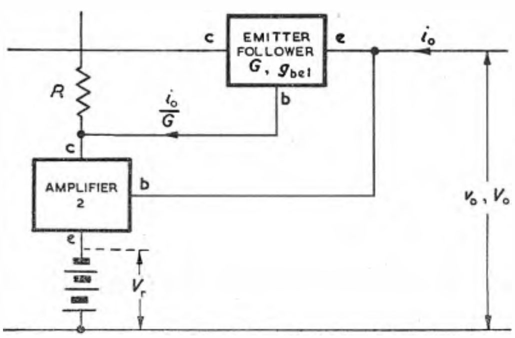


Fig. 3. The application of d.c. feedback

can be virtually eliminated by supplying the feedback amplifier from a separate stabilized line.

In detail, the output resistance may be calculated as before by considering that a small low frequency current i_o flowing in the output circuit causes a voltage v_o to appear across the output terminals.

If the amplifier consists of a single grounded emitter transistor of base-collector transconductance g_{bc2} which has an a.c. load R comprising a physical load and the output resistance of the transistor, then assuming that the input current to the amplifier is a negligible fraction of i_o

$$[g_{bc2}v_{be2} - (i_o/G)]\, R = -v_o + v_{be1}$$

$$v_{be2} = v_o - g_{be2}v_{be2}r$$

$$\therefore\ v_{be2} = \frac{v_o}{1 + g_{be2}r}$$

$$\frac{g_{bc2}Rv_o}{1 + g_{be2}r} - (i_oR/G) = -v_o + (i_o/g_{be1})$$

so that

$$r_o = (v_o/i_o) = \frac{1/g_{be1} + (R/G)}{1 + (g_{bc2}R/(1 + g_{be2}r))} \quad \dots\dots\dots \quad (4)$$

In a typical case at 100mA with a single transistor amplifier and a compound emitter-follower of two transistors

$(1/g_{be1}) = 1\Omega$ $\quad R = 10\text{k}\Omega$ $\quad G = 2\,500$
$g_{be2} = 40\text{mA/V}$ $\quad g_{bc2} = 40\text{mA/V}$ $\quad r = 2\Omega$
$r_o = 0{\cdot}013\Omega$

Thus the effect of applying feedback is to reduce the output resistance by a factor of several hundred. The d.c. output voltage is given by

$$V_o = V_r + V_{be2}$$

Variations in the value of V_{be1} with temperature cause only a small variation in V_{be2} due to the feedback loop, and this is negligible compared with the 2·5mV/°C by which V_{be2} varies on its own account.

This variation of output voltage with temperature can be minimized by using a long-tailed pair as the amplifier (Fig. 4). The output voltage is now given by

$$V_o = V_r + V_{be2}$$

and since temperature variations in V_{be3} and V_{be4} are equal they will be balanced out, but the effect of I_{co} of X_4 flowing in the source resistance of the reference must now be considered.

The terms g_{bc2} and g_{be2} of equation (4) can now be derived in terms of the circuit of Fig. 4.

g_{bc2} *is* the ratio of the incremental collector current of X_3 to the incremental voltage between b_3 and b_4

$$i_{c3} = \frac{-\alpha_3}{r_{e3} + (r_{b3}/(1+\alpha_3'))} \cdot v_{be3}$$

$$i_{e3} = \frac{-1}{r_{e3} + (r_{b3}/(1+\alpha_3'))} \cdot v_{be3}$$

$$i_{e4} = \frac{-1}{r_{e4} + (r_{b4}/(1+\alpha_4'))} \cdot v_{be4}$$

$$i_{e3} + i_{e4} = 0$$

$$\frac{v_{be3}}{r_{e3} + (r_{b3}/(1+\alpha_3'))} + \frac{v_{be4}}{r_{e4} + (r_{b4}/(1+\alpha_4'))} = 0$$

$$v_{be4} = -v_{be3}\left\{\frac{r_{e4} + (r_{b4}/(1+\alpha_4'))}{r_{e3} + (r_{b3}/(1+\alpha_3'))}\right\}$$

$$v_{b3b4} = v_{be3} - v_{be4}$$

$$= v_{be3}\left\{1 + \frac{r_{e4} + (r_{b4}/(1+\alpha_4'))}{r_{e3} + (r_{b3}/(1+\alpha_3'))}\right\}$$

so that the term g_{bc2} of equation (4), being equal to

$$-i_{c3}/v_{b3b4}$$

becomes

$$g_{bc2} = \frac{\alpha_3}{r_{e3} + r_{e4} + \dfrac{r_{b3}}{1 + \alpha_3'} + \dfrac{r_{b4}}{1 + \alpha_4'}} \quad \dots\dots\dots\dots \quad (5)$$

which is approximately half the value compared with a single transistor. The value of $r_{e3} + r_{e4}$ in the long tailed pair may be calculated by considering

$$r_{e3} = (25/I_{e3}) \qquad r_{e4} = (25/I_{e4})$$

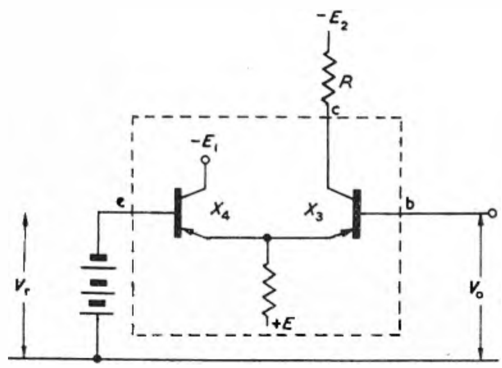


Fig. 4. Use of 'long-tailed' pair to minimize variation of output voltage with temperature

and $I_{e3} + I_{e4} = I$ which is constant.

$$\therefore r_{e3} + r_{e4} = 25\,(1/I_{e3} + 1/I_{e4})$$

$$= \frac{25\,(I_{e3} + I_{e4})}{\frac{1}{4}[(I_{e3} + I_{e4})^2 - (I_{e3} - I_{e4})^2]}$$

$$= \frac{100\,I}{I^2 - \delta^2} \simeq (100/I)\,(1 + (\delta^2/I^2)) \ldots\ldots\ldots \quad (6)$$

where δ is the out-of-balance current which is small compared with I.

g_{be2} is the ratio of the base current of X_4 to the voltage between b_3 and b_4

$$i_{b4} = (1/1 + \alpha_4')\,.\,i_{e4} = \frac{-1}{r_{b4} + (1 + \alpha_4')r_{e4}}\; v_{be4}$$

$$v_{be3} = v_{be4}\,\frac{r_{e3} + (r_{b3}/(1 + \alpha_3'))}{r_{e4} \div (r_{b4}/(1 \div \alpha_4'))}$$

$$v_{b3b4} = v_{be3} - v_{be4}$$

$$= -v_{be4}\left\{1 + \frac{r_{e3} \div (r_{b3}/(1 \div \alpha_3'))}{r_{e4} + (r_{b4}/(1 + \alpha_4'))}\right\}$$

$$g_{be2} = (-i_{b4}/v_{b3b4}) = \frac{1}{(1 + \alpha_4')\left(r_{e3} + r_{e4} + \frac{r_{b3}}{1 + \alpha_3'} + \frac{r_{b4}}{1 + \alpha_4'}\right)} \ldots\ldots\ldots \quad (7)$$

which is a reduction of approximately $2\alpha'$ compared with a single transistor.

Equation (4) can now be rewritten in terms of the long-tailed pair by substitution in equation (5), (6) and (7)

$$r_o = (1/g_{be1}) + (R/G) \div$$

$$1 + \frac{R\alpha_3}{(100/I)\,(1 + (\delta^2/I^2)) + \frac{r_{b3}}{1 + \alpha_3'} + \frac{r_{b4} + r}{1 + \alpha_4'}}$$

In a typical case at 100mA with $r = 2\Omega$

$$I = 1{\cdot}5\text{mA} \qquad \delta = 0{\cdot}5\text{mA}$$

$$1/g_{be1} + (R/G) = 5\Omega \text{ as before}$$

$$1 + \frac{R}{100/I\,(1 + (\delta^2/I^2)) + \frac{r_{b3}}{1 \div \alpha_3'} + \frac{r_{b4} + r}{1 + \alpha_4'}} = 1 + \frac{10^4}{75 + 8 + 8}$$

$= \text{III}$ so that $r_\alpha = 0{\cdot}047\Omega$.

It can be seen that using this system of d.c. feedback, the d.c. output resistance may be made extremely low. The use of the long-tailed pair does not have a very great effect on the output resistance, but it does make the output resistance less dependent on the internal resistance of the reference source, and the output voltage less dependent on temperature provided that the internal resistance of the reference source is low.

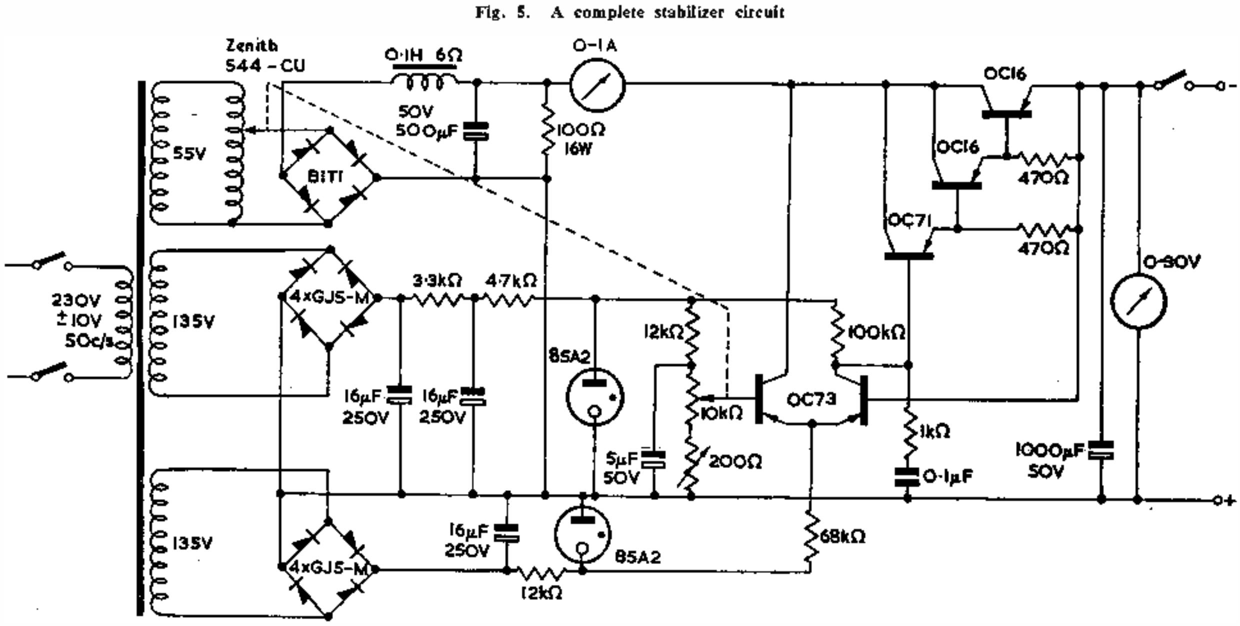
A Practical Circuit (Fig. 5)

Transistor Dissipation

The dissipation in the series transistors provides a limit to the voltage which may be dropped across them from the unstabilized supply. Hence in a unit supplying a variable output voltage up to 30V it is not feasible to provide an unstabilized supply which is not varied with the output voltage required. Also, in order that the series transistors are not bottomed at full loads and are not subjected to an excessive voltage on no load, the output resistance of the unstabilized supply should be not greater than a few ohms. In order to fulfil these requirements, the supply used a choke input filter with 100Ω load, and a low resistance variac, ganged to the reference potentiometer was used to vary the unstabilized supply. The mains input voltage is also required to be held within 10V of its nominal value.

The output resistance of the supply used was approximately 9Ω and in order to prevent the series transistors bottoming on the peaks of the 100c/s ripple at full load,

Fig. 5. A complete stabilizer circuit



and allowing for mains variations of ± 10V, the minimum voltage across the transistors was limited to 3V.

The power dissipated in the compound emitter-follower never exceeded 6W and most of this was dissipated in one of the OC16's. This transistor is therefore mounted directly on a 16 s.w.g. copper plate approximately 6in square with a matt black finish, and mounted in a position assuring adequate ventilation. The thermal resistance of the heat sink in this position was 2·6°C/W, allowing a maximum of 6·8W dissipation in the OC16 at an ambient temperature of 45°C.

Choice of Reference Source

The only stable reference source readily available was the 85A2 neon reference tube. It was decided that this, in conjunction with a potentiometer network, would be satisfactory since this allowed the output voltage to be varied by means of the reference voltage rather than by adjusting the level of the feedback voltage which involves variations of loop gain with output voltage.

The Long-tailed Pair

The coupled emitters were required to be run from a stable positive line rather than the earth line, so that control could be maintained right down to zero output voltage. This was achieved by providing a positive voltage stabilized with a second 85A2. In order to achieve a high gain in the amplifier and to make the output voltage independent of the input, the 100kΩ collector load was connected to the -85V reference line.

Choice of Emitter-follower Configuration

The maximum output current required was 1A, so that to obtain adequate current gain it was decided that a compound circuit using three transistors would be necessary. These were chosen to be OC16, OC16, OC71.

The centre one was chosen to be an OC16 since, although the OC72 would meet the current requirement, it would not meet the dissipation requirement with the collector voltages expected and at the maximum specified ambient temperature of 45°C. The uses of two parallel transistors in this connexion was not considered suitable since the addition of series resistance to ensure adequate current sharing could seriously affect the output resistance of the unit. The current gain of this compound emitter-follower was approximately 10^5.

Frequency Response

Due to the fall of α' with frequency, the output impedance of the device rises with the frequency of the load current variations. In addition, phase shift in the four-transistor loop causes oscillation to occur at a frequency of approximately 10kc/s. To counter this, a resistor of 1kΩ in series with a capacitor of 0·1μF shunts the output of the long-tailed pair to reduce the gain of the amplifier at the frequencies at which the phase shift causes regeneration.

Performance Figures

The average output resistance of the unit over the current range 0 to 1A was less than 0·04Ω at d.c. and less than 0·2Ω at all frequencies less than 100kc/s.

The change in output voltage due to a 10V change in mains input was 0·2 per cent, most of this being due to variations in the reference and the 100c/s ripple was less than 1mV under all conditions.

The variation of output voltage with temperature over the range 20 to 50°C was approximately 200mV. Most of this was due to I_{co} flowing in the comparatively high resistance of the reference potentiometer, the variation between output and reference point being less than 5mV.

Acknowledgment

The authors wish to thank the Manager of Mullard Research Laboratories and the Directors of Mullard Ltd. for permission to publish this article. They are indebted to their many colleagues and particularly to Mr. L. P. Morgan for helpful discussions, and to Mr. K. C. Johnson of Ferranti Ltd. for useful criticism.

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A New Ultrasonic Cleaning Bath

The L.276, a new type of ultrasonic cleaning bath for small components and precision piece parts, is announced by Mullard Ltd. as being in full production.

The Company's experience in the development and application of ultrasonic techniques has shown a specific need for a compact, hand-fed cleaning bath to supplement conveyorized equipment for the rapid and effective cleaning of small articles. Much of the design of this new bath is directly related to extensive field investigations undertaken to determine the precise requirements of many industries.

The outstanding performance of the L.276 is attributable largely to a new design of low frequency radial transducer, developed specifically for the equipment. This transducer has two important advantages, both of which result in improved cleaning efficiency and shorter processing times. First, it effectively focuses the cavitation intensity in the centre of the cleaning fluid container, where the component would be situated. Second, the cavitation effect is directed equally to all sides of the component simultaneously: therefore, unless exceptional requirements call for multi-stage processing, the entire surface area of the article can be cleaned with one dip.

Yet another advantage of the equipment is that, due to its special construction, hot solvents may be used direct. The fluid container (a standard Pyrex glass beaker) and the transducer are contained in a water jacket. The water serves not only to cool the transducer, but also to couple the ultrasonic energy to the cleaning fluid.

The beaker is easily removable—it simply stands in position in the centre of the transducer annulus; multi-stage processing with a variety of cleaning agents when exceptionally high cleanliness is required can therefore be carried out by simple substitution of beakers. Furthermore, since only the beaker is in direct contact with the fluid, no problems arise with the cleaning of the equipment.

Capacity of the beaker is 250ml, and its dimensions are approximately 2¾in diameter by 4½in deep.

The radial transducer incorporated is a low-frequency type operating at 20kc/s. The choice of frequency (low as opposed to high) is dictated largely by the nature of the contaminations with which the cleaning bath is intended to deal.

Very briefly, the cleaning effect obtained by applying ultrasonic energy to the cleaning fluid is due to two factors. One is the acceleration given to the particles of the liquid, which then bombard the surface of the component. The other is the subsidiary pressure effects of cavitation which occur in the liquid. This results in a scrubbing action against any exposed surfaces which gives a far more thorough and rapid cleaning than is attainable by conventional methods. At high ultrasonic frequencies (say 1Mc/s) the significant factor is the acceleration and high beam pressures—the extent of cavitation is very small. At low frequencies the cavitation effect is great and the acceleration low.

For the removal of tenacious deposits, the intense cavitation which occurs at low frequencies, and which is largely responsible for the scrubbing action, is extremely effective. Lightly-held contamination is best removed by the application of high frequency energy, the impurities being flushed away by the high forward beam pressures which occur in the absence of cavitation.

The Modulation of Travelling-Wave Tubes

By G. F. Steele*, B.Sc.

The simple theory of travelling-wave tube modulation is discussed and some expressions derived for the cases of phase and amplitude modulation. Details are given of the experimental procedure used to measure phase shift, and the sideband power resulting from a 60Mc/s modulation.

THE travelling-wave tube is a broadband amplifier of microwave frequencies and its use in communications and television links is well known. Before discussing the actual mechanism of modulation, the principles underlying the use of a travelling wave tube as a modulator will be briefly restated. The use of a travelling-wave tube as a phase modulator was originally discussed by W. H. Bray[1].

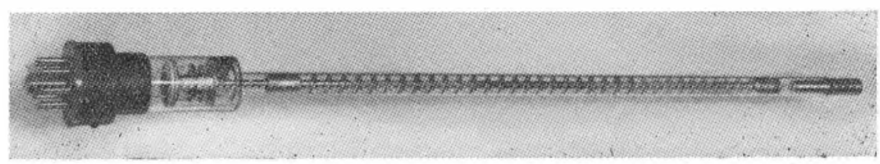
If a signal of about 15Mc/s bandwidth is to be transmitted at microwave frequencies, the spectrum of such information will be a carrier at a microwave frequency with the information occupying sidebands on either side of it. As at present used, the travelling-wave tube is not directly modulated with the information, but is modulated by a sub-carrier which carries the information. In this sense, the travelling-wave tube is used as a frequency shifter rather than a modulator.

Suppose a travelling-wave tube is modulated by a signal of frequency f_1 (60Mc/s) which also carries the information, designated by Δf. The way in which this signal may be applied to the tube is explained later. If the tube also has a microwave signal of frequency f_o injected at the normal input, the output of the tube will now be $f_o \pm n\ (f_1 \pm \Delta f)$, where n is an integer. The amplitude of the components will clearly depend upon the kind of modulation used.

If one of the sidebands is selected by means of a band-pass filter one now has the required microwave carrier and information. The system is shown in diagrammatic form in Fig. 1.

The final signal is a microwave carrier at a frequency of $f_o + f_1$ which has the information occupying a bandwidth of Δf.

At a repeater station, the microwave frequency can easily be shifted by injecting a c.w. modulating signal.



*MS*1 200. *Medium Power Tube* 3 600-4 200*Mc/s*

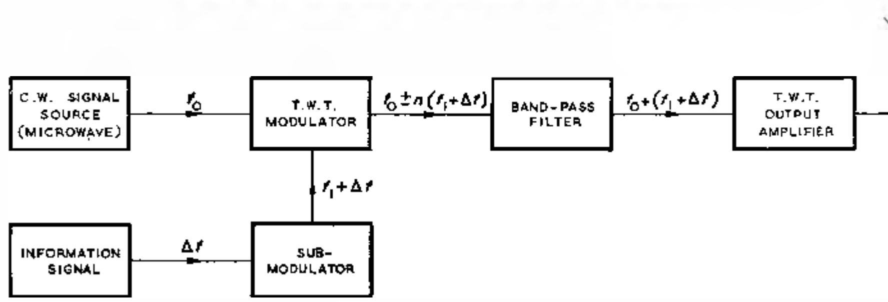


Fig. 1. Use of travelling-wave tube as a modulator

Travelling-Wave Tube Gain Equations and Modulation Theory

The output voltage, E, of a travelling-wave tube may be represented by an equation of the form

$$E = E_o \exp(j\omega t - \Gamma z),$$

where Γ is the propagation constant. The symbols and parameters used in this article are those given by Pierce[2]. Now $\Gamma_i = j\beta_e + \beta_e C\delta_i$ $(i = 1, \ldots .4)$ where C is the gain parameter and β_e is the phase constant corresponding to the electron velocity. Γ and δ have four values corresponding to the four waves propagated. Only the forward increasing wave is of interest, the one which gives rise to the gain of the travelling-wave tube, where δ is given by

$$\delta = (\sqrt{3} - j)/2$$

Hence $\Gamma = -j\beta_e + \beta_e C(\sqrt{3} - j)/2$

Therefore $E = E_o \exp[j\omega t + \beta_e C(\sqrt{3}/2)z - j\beta_e(1 + C/2)z]$

The real part of this is

$$E_o \cos[\omega t - \beta_e(1 + C/2)z] \exp[\beta_e C(\sqrt{3}/2)z$$

If z_1 is the effective axial length of the helix, it is seen from the above expression that the phase and amplitude of the output are determined by $\beta_e(1 + C/2)z_1$ and $\exp(\sqrt{3}\beta_e C z_1/2)$ respectively.

In addition there are two relations defining β_e and C, viz:

$$\beta_e = \omega/u_o$$

$$C \propto (I_o/V_o)^{1/3}$$

where u_o is the beam velocity, which is dependent on the helix voltage, V_o, and I_o is the beam current.

The expressions for phase and amplitude will now be used to determine the performance of travelling-wave tubes as modulators.

Phase Modulation

So far travelling-wave tubes have been phase modulated in the systems described in the introduction and this method will be considered first.

The phase, ϕ, of the output relative to the input is given by

$$\phi = \beta_e(1 + C/2)z_1$$
$$= (\omega/u_o)(1 + C/2)z_1$$

But $u_o = kV_o^{1/2}$

Therefore if $C \ll 1$,

$$\delta\phi/\partial V_o = -\omega z_1/2kV_o^{3/2}$$
$$= -\omega z_1/2u_oV_o$$

The magnitude of the output is also dependent upon V_o, but changes in magnitude are small and may be neglected,

* Mullard Ltd.

for small changes in V_0. These conditions are fulfilled in practice.

It is clear from the expression derived for phase shift that a tube operating from a large helix voltage will give less phase shift per volt that one operating with a lower voltage. To illustrate this, experiments were performed on two tubes which operate at different helix voltages; a brief description of these tubes is given below.

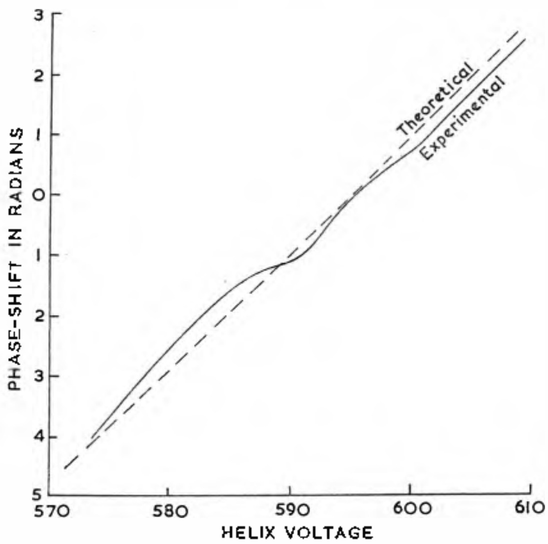


Fig. 2. VX8168 phase shift with helix volts

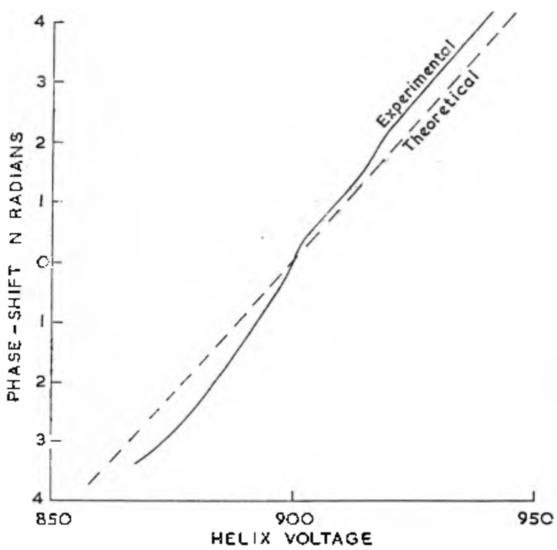


Fig. 3. Phase shift/helix voltage for MS1200

TRAVELLING-WAVE TUBES USED FOR MODULATION EXPERIMENTS

The lower voltage tube is the VX8168 which is a low noise tube giving a gain of 15dB and a saturation power output of 2mW. It is doubtful whether it would be useful in practice for modulation purposes as its saturation output power is low, but it provides a useful check on the theory at low helix potentials.

A tube with a somewhat higher operating voltage is an experimental version of the MS1200 (medium power tube) which will deliver 200mW under saturation conditions at a gain of 25dB.

Both tubes work in the 4kMc/s communications band but neither was designed specifically for modulation purposes.

EXPERIMENTAL RESULTS

Graphs of phase shift versus helix voltage are shown in Figs. 2 and 3 for the low and high voltage tubes respectively.

If the phase shift for a particular change in helix voltage

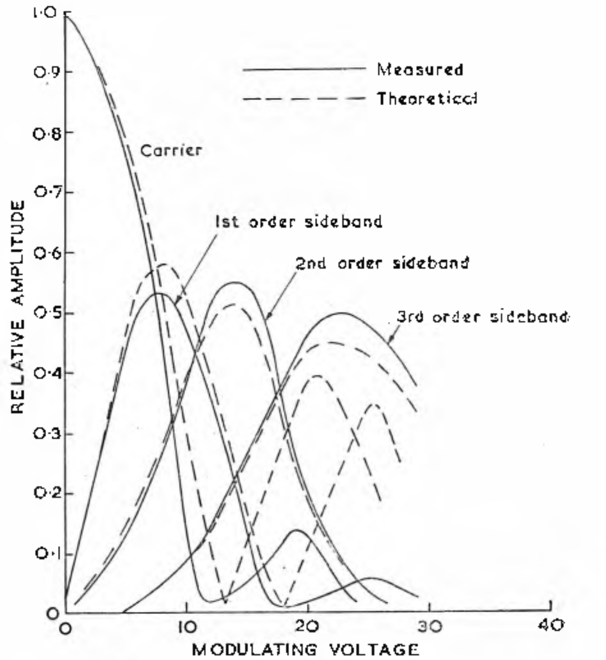


Fig. 4. Phase modulation of VX8168

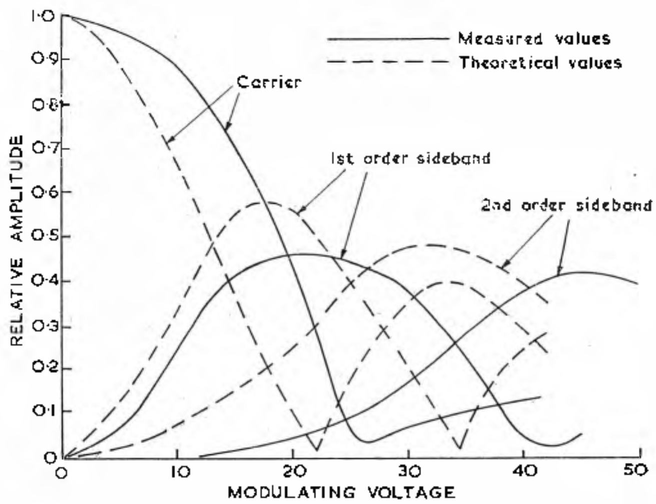


Fig. 5. Phase modulation of MS1200

is known, then the expected result, if an oscillatory signal were applied to the helix, is easily calculated. A multiplicity of sidebands is produced in a phase modulated wave and in practice the modulating voltage is adjusted so that the modulation index is equal to 1·8, where the first order sideband reaches its maximum value. This sideband is selected by means of a band-pass filter.

The results that were obtained by applying a 60Mc/s signal to the helices of the two tubes are shown in Figs. 4 and 5. It will be seen that the applied voltage is higher than the theoretical one for a given sideband maximum, which is due to the fact that the voltage applied to the base pins of the tube is not the same as the actual helix to cathode voltage on account of the long leads used in the tube construction.

Details of the experimental procedure used in modulating the tubes and of the measurement of sideband power are given in the Appendices.

Amplitude Modulation

One of the problems attached to the phase modulation of travelling-wave tubes, is the fairly high modulating voltage necessary to obtain maximum power in the first order sideband. In particular, it becomes increasingly difficult as the carrier frequency increases, since travelling-wave tube design theory precludes the use of very low voltage helices at the higher microwave bands. In addition there is a move to make tubes shorter and so reduce the effective length, z_1, which again reduces the phase shift per volt. It would seem therefore to be worthwhile examining the possibility of amplitude modulation.

Amplitude modulation is mentioned in the literature as being possible, though no published results are available. It has been suggested that, by the use of a control grid very near the cathode, amplitude modulation might be possible at extremely low voltages.

Amplitude of the output voltage

Consider the expression, $\exp(\beta_e C\sqrt{3}z/2)$, which determines the magnitude of the voltage output of the tube.

$$C = (I_o/8V_o)^{1/3}K^{1/3}$$

where K is an impedance parameter,

and $$I_o = k_1 V_A^{3/2}$$

where V_A is the voltage between the cathode and anode of the electron gun, and k_1 is a constant.

Therefore

$$C = (k_1 V_A^{3/2}/8V_o)^{1/3}K^{1/3}$$

Hence if V_o is a constant,

$$C = kV_A^{1/2}$$

where $$k = (k_1 K/8V_o)^{1/3}$$

Hence one may write for the amplitude of the output of the tube,

$$\exp[\beta_e C(\sqrt{3}/2)z_1] = \exp[\beta_e k V_A^{1/2}(\sqrt{3}/2)z_1]$$
$$= \exp[FV_A^{1/2}]$$

since β_e and z_1 are constant.

$$F = \beta_e k(\sqrt{3}/2)z_1 \qquad \text{(a constant)}$$

Suppose now a modulating voltage is applied to the first anode such that

$$V_A = V_a + v\cos pt$$

The magnitude of the output is now equal to

$$\exp[F(V_a + v\cos pt)^{1/2}] \simeq \exp[FV_a^{1/2} + (Fv/2V_a^{1/2})\cos pt]$$

Put $$a = Fv/2V_a^{1/2}$$

Then, as F will be shown to be small, a is small, hence

$$\exp[Fv/2V_a^{1/2})\cos pt] = \exp(a\cos pt)$$
$$= 1 + a\cos pt + a^2(1 + \cos 2pt)/4 + \ldots\ldots$$
$$\simeq 1 + a^2/4 + a\cos pt + (a^2\cos 2pt)/4$$

Thus the output of the tube has a magnitude determined by

$$\exp(FV_a^{1/2})\left(1 + \frac{F^2v^2}{16V_a} + \frac{Fv}{2V_a^{1/2}}\cos pt + \frac{F^2v^2}{16V_a}\cos 2pt\right)$$

The actual output of the tube in this expression multiplied by

$E_o\cos(\omega t + \phi)$ viz:

(a) $E_o\exp(FV_a^{1/2})\,(1 + F^2v^2/16V_a)\cos(\omega t + \phi)$

(b) $+\ \frac{1}{2}E_o\exp(FV_a^{1/2})\,(Fv/2V_a^{1/2})\,(\cos[(\omega + p)t + \phi] + \cos[(\omega - p)t + \phi])$

(c) $+\ \frac{1}{2}E_o\exp(FV_a^{1/2})\,(F^2v^2/16V_a)\,(\cos[(\omega + 2p)t + \phi] + \cos[(\omega - 2p)t + \phi])$

(a) gives the carrier wave,

(b) gives the first order sideband terms,

(c) gives the second order sideband terms,

For numerical calculations, F may be determined from the performance of the tube as an amplifier.

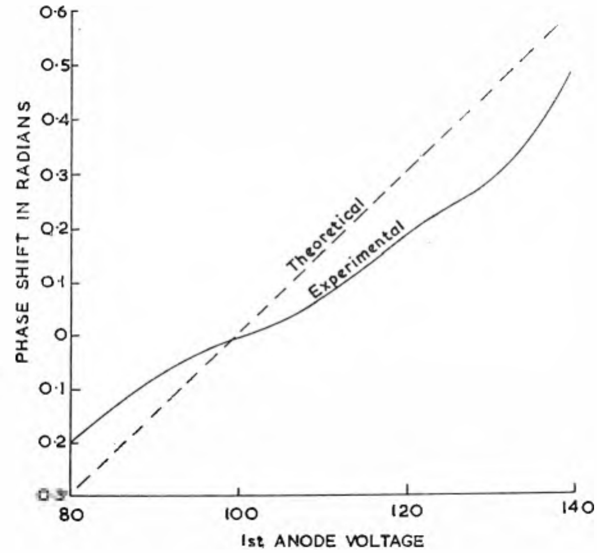


Fig. 6. Phase shift-first anode voltage (changes in beam current) for VX8168

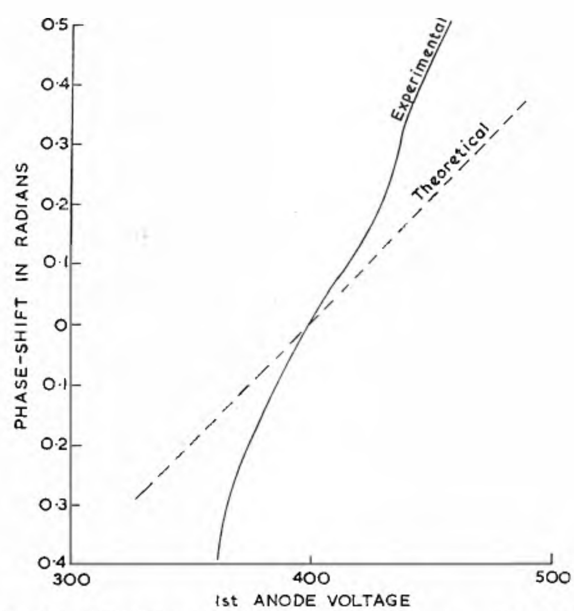


Fig. 7. Phase shift-first anode voltage (changes in beam current) for MS1200

The gain of the tube in decibels is given by an expression

$$G = A + BCN$$

where A is a term which expresses the initial loss of the tube.

Also the voltage ratio of the gain $= \exp(FV_A^{1/2})$

Hence, if the measured gain of the tube is G_1, then $G_1 - A$ expressed as a voltage ratio is $\exp(FV_A^{1/2})$, and hence F may be evaluted.

For the MS1200, A is -12 and $F \simeq 0{\cdot}25$.

Phase Shift

Before comparing the theoretical modulation with the figures obtained in practice one must look again at the expression for phase shift,

$$\phi = \beta_e(1 + C/2)z_1$$

where it will be noticed that C also occurs.

$$C = kV_A^{1/2}$$

hence $\phi = \beta_e(1 + kV_A^{1/2}/2)z_1$

therefore $\partial\phi/\partial V_A = \beta_e k z_1/4V_A^{1/2}$

It has also been shown that

$$\beta_e k z_1 \sqrt{3}/2 = F$$

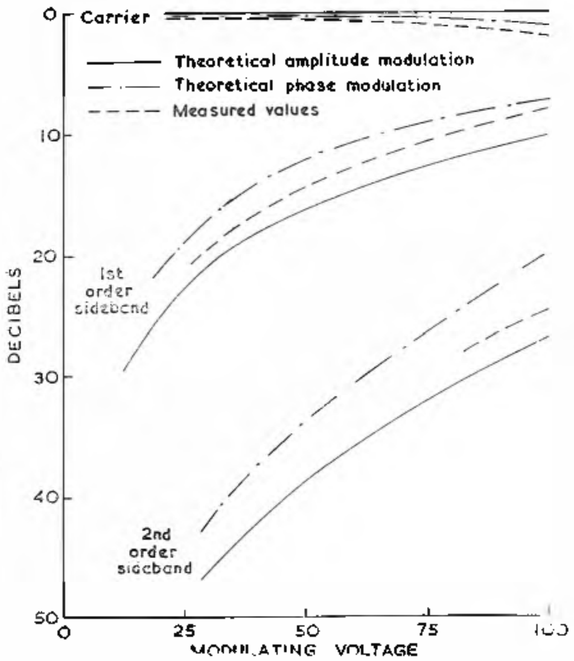


Fig. 8. Cathode-first anode modulation of MS1 200

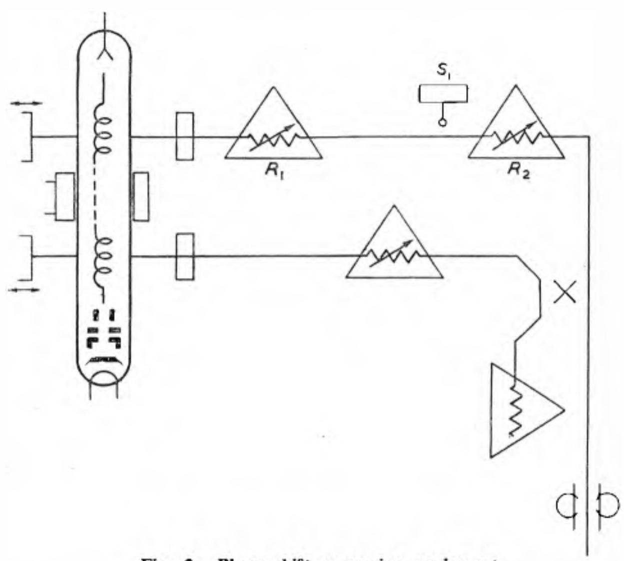


Fig. 9. Phase shift measuring equipment

therefore $\beta_e k z_1 = 2F/\sqrt{3}$

therefore $\partial\phi/\partial V_A = F/2\sqrt{3}V_A^{1/2}$

The experimental results for the two tubes are shown in Figs. 6 and 7.

The agreement between theoretical and practical results seems moderate in the case of a tube with low C (VX8168) but poor in the case of a tube with higher C (MS1200). It may be that in the latter case, the changes in the beam profile with changes in the first-anode voltage exert considerable influence. Concurrently with the work described here, the phase shift produced by d.c. changes in beam current has been investigated by Beam and Blattner[3] whose results for the case of a tube with low C agree with the observations made here.

From the experimental curve of phase shift versus first-anode voltage it has been possible to estimate the degree of phase modulation that would take place if the amplitude remained constant. For the MS1200, Fig. 8 shows:

(a) The theoretical amplitude modulation in the absence of phase shift.

(b) The theoretical phase modulation in the absence of amplitude changes.

(c) The actual modulation obtained by applying a signal between cathode and first-anode.

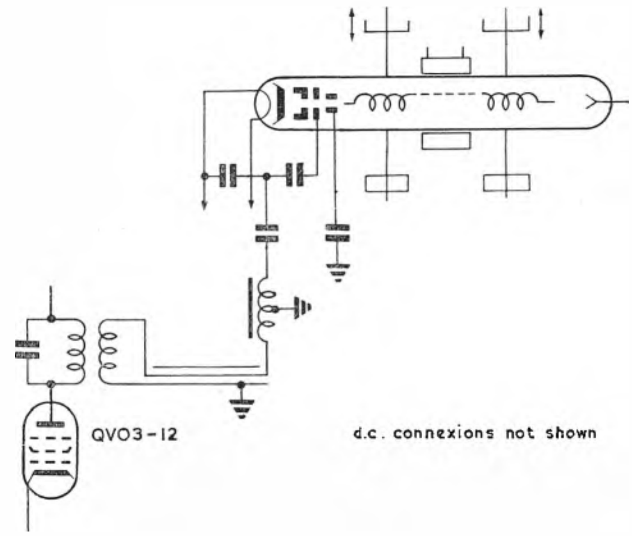


Fig. 10. Circuit for phase modulation

It must also be remembered that there is, inevitably, coupling between the helix and first-anode resulting in further distortion. This kind of distortion is quite small in the case of the MS1200, but with the VX8168, because of the construction of the low noise gun, attempts to apply a signal between cathode and first anode were frustrated by coupling on the helix.

Conclusions

Phase modulation of travelling-wave tubes appears to be much more satisfactory than amplitude modulation, with conventional tubes. The phase shift which occurs with changes in beam current appears to depend upon the type of tube used, and is quite severe in the case of the medium power tube used here. Hitherto however, manufacturers of equipment have used as modulators travelling-wave tubes that have not been designed specifically for that purpose. In consequence of this the tubes may not be as efficient as possible.

In view of the difficulty experienced in trying to amplitude modulate the VX8168 more attention must be given to gun design if a tube is required for modulation purposes. It is difficult to apply a 60Mc/s signal at present; a 200Mc/s signal which may be desirable in some circumstances would present an even greater problem.

Appendix

(1) Measurement of Phase Shift

The method used was a conventional one and can be understood from Fig. 9.

From the signal source, power is fed to both input and output of the travelling-wave tube by means of a directional coupler. On the output side of the tube a standing wave is produced due to the combination of the output of the tube and the signal fed to it as described above. If a standing wave indicator is included as shown in the diagram, the changes in null point of the standing wave pattern enable the change of phase of the output to be determined.

Suppose the wave from the output of tube incident upon the standing wave indicator is

$$E^{+} = E_0 e^{(j\omega t + j\gamma z + \phi)}$$

while the signal direct from the source is

$$E^{-} = E_0 e^{j(\omega t - \gamma z)}$$

In practice these waves can be made of equal amplitude by adjusting the attenuators A_1 and A_2.

The resultant $E_r = E^{+} + E^{-}$

$$= E_0 e^{j\omega t}\left(e^{-j\gamma z} + e^{(j\gamma z + \phi)}\right)$$

At the point where $E_r = 0$

$$e^{-j\gamma z} = -e^{j(\gamma z + \phi)}$$

Therefore

$$e^{-2j\gamma z} = -e^{j\phi}$$

Therefore

$$\phi + (2n + 1)\pi = -2\gamma z$$

Therefore

$$\delta\phi = -2\gamma\delta z = -4\pi\delta z/\lambda_g$$

Hence the phase shift can be found in terms of changes in the position of the null point.

Having adjusted the attenuators A_1 and A_2 to give a true zero it was found desirable to leave them unaltered during a series of measurements even though the output of the tube may vary. The reason for this was that a change of attenuation itself produced small phase shifts.

(2) Application of the 60Mc/s Signal to the Tubes

To provide the fairly high voltages required for the tests, a narrowband amplifier was built using a QVO3-12 as the output stage.

For phase modulation, the voltage must be applied between helix and cathode. The helix-earth capacitance is high (the various interelectrode capacitances are listed below) and for this reason the helix was operated at earth potential with respect to a.c. and the signal applied between cathode and earth. To prevent unwanted amplitude modulation the cathode and first anode were connected together as far as r.f. is concerned. The type of circuit used for each tube is shown in Fig. 10.

In the case of amplitude modulation the helix and cathode were both earthed with respect to r.f. so that no phase modulation due to cathode-helix voltage changes could take place. With this precaution the signal was applied between first anode and earth. The type of circuit used is shown in Fig. 11.

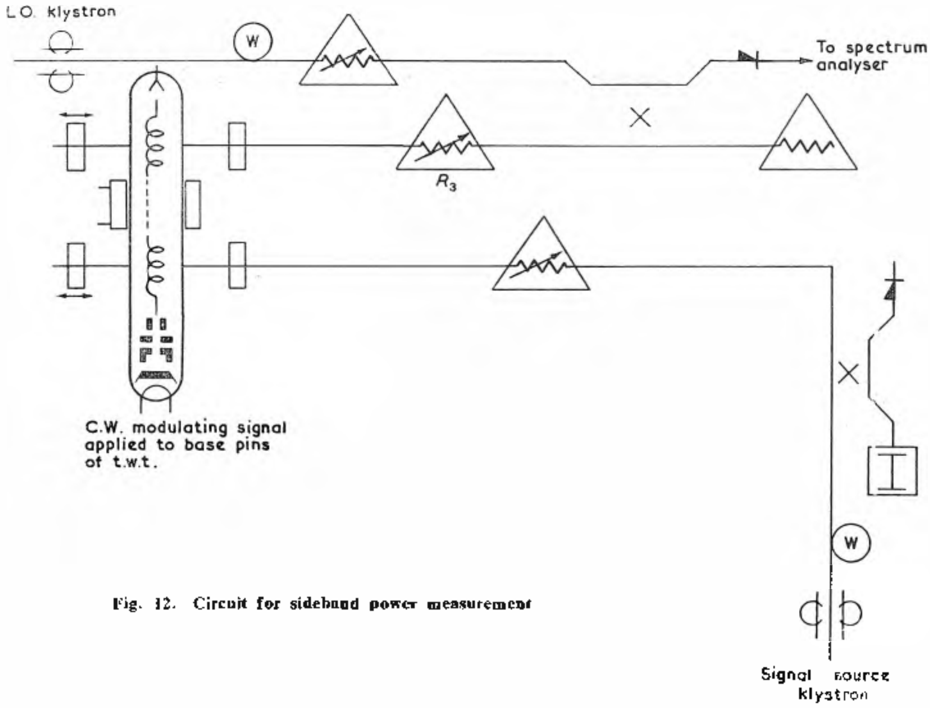
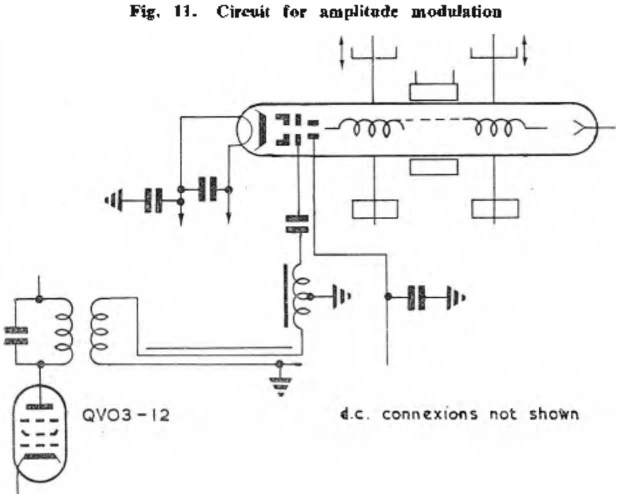


Fig. 12. Circuit for sideband power measurement

The various interelectrode capacitances are:

	VX8168	MS1200
Helix-earth	28pF	37pF
First anode-helix	22pF	8pF
First anode-cathode	3pF	4pF
Helix-cathode	2pF	1pF

Fig. 11. Circuit for amplitude modulation



(3) Measurement of Sideband Power

The lack of suitable band-pass filters or appropriate coupling cavity, for selection of the sidebands, led to the use of a spectrum analyser for measurement of sideband power. From a knowledge of the intermediate frequency of the analyser, the signal source frequency and the local oscillator frequency, any particular sideband of interest could be selected. Provided that the mixer crystal current was kept constant and the gain of the analyser fixed, the response of the analyser was found to be substantially dependent upon the power of the microwave signal under investigation. At each frequency of interest, the response of the spectrum analyser was calibrated against a precision attenuator (A_3 in the figure). The experimental circuit is shown in Fig. 12.

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Approximating Non-linear Functions

by Shunt Loading Tapped Potentiometers in Analogue Computing Machines

By D. W. C. Shen*, Ph.D.

Simulation of non-linear functions in the aerodynamic section of an analogue computer can be achieved by linear interpolation techniques. Diode function generators used for this purpose present several items for consideration by both designers and operators. These items are careful selection of valves having contact potentials within admissible range, calibration procedure for repeatability of results and the limitation of the size of the problem to be simulated by the amount of equipment, particularly amplifiers, available in a given machine. As an alternative, a shunt potentiometer type function generator has a number of advantages, i.e., accuracy, flexibility, simplicity in operation and repeatability. Accuracy and repeatability are due to certain averaging properties of the resistance networks.

The design of the shunting network is greatly simplified by a method of network tearing. A non-monotonic function can be divided both mathematically and electrically into distinct monotonic sections and the calculations of suitable shunting resistors, end resistors and pull-in resistors reduce to a simple routine. The relation between minimum allowable winding resistance, maximum functional slope and input impedance sets the design criterion.

The application of tapped potentiometers to simulate aerodynamic force and torque equations is briefly described. To extend the utility of tapped potentiometers, polynomial function generators and clamped potentiometers can corporate the shunting techniques for special purposes.

RECENT publications[1,2,3] on the electronic analogue computers have demonstrated the effectiveness of such machines in solving dynamic problems of complex engineering systems. Large machines such as Tridac[4] are extremely useful in connexion with research on automatic control of flight. The aerodynamic equations for aircraft and missiles are mostly non-linear. After the aerodynamic data have been smoothed and converted into a tractable form, it is necessary to represent them by function generators in an analogue machine. Fortunately, most non-linear data may in most cases be adequately represented by the first few terms of a power series. For some cases simple forms, which ignore some cross-coupling terms can be used. By incorporating function generators in the analogue computers, the scope of such machines can be widely extended to solve problems involving non-linear equations.

One type of function generator known as a photo-former[5] requires special apparatus. Its operation depends upon the trace of a profile of a mask placed on the surface of a cathode-ray tube by a light spot. The profile corresponds to the function to be generated. To simulate aerodynamic equations by a number of photo-formers is both expensive and complicated. Recognizing that aerodynamic data are not generally known with great accuracy, consequently extremely accurate simulation is not generally justified. Function generators based on linear interpolation techniques are much preferred in such applications. It may be mentioned that recent work on linearization and quasi-linearization techniques applied to non-linear control[6,7] systems have yielded fruitful results. Burt and Lange[8] described a technique in which diode units are used to approximate non-linear functions by linear interpolation. It was shown that this technique can be extended to cover a wide class of functions instead of restricting to monotonic increasing functions as described in early literature. However, using biased diode units as function generators requires normally two high gain drift stabilized d.c. amplifiers for each generator. The amount of equipment needed for analoging aerodynamic equations, particularly amplifiers, may exceed the capacity of the machine. Furthermore, the operation and stability of diode function generators are effected by diode characteristics such as heater voltage, contact potential, diode resistance etc. Usually the repeatability of results is not satisfactory and this requires a calibration procedure which must be designed to reduce the effort to a minimum in order to achieve ease of operation.

This article shows that non-linear functions can be achieved in precision potentiometers by shunt loading several points of the resistance elements. By providing the potentiometer winding with several taps, a variety of non-linear curves can be approximated by attaching appropriate shunt resistors. The principal features of shunt loading techniques are their flexibility, simplicity and repeatability. By a method of network tearing, the design

LIST OF PRINCIPAL SYMBOLS

$f(x) =$	an arbitrary function
$P_n(x) =$	n^{th} linear line segment approximating $f(x)$
$\xi_a, \xi_b =$	errors due to linear interpolation
$R_w =$	winding resistance of an unshunted tapped potentiometer
$R_s =$	shunt resistance of a potentiometer winding segment
$R_e =$	end resistance
$R_p =$	pull-in resistance
$Z_i =$	input impedance of the shunted tapped potentiometer
$b =$	number of monotonic sections in a non-linear function
$V_j =$	voltage at node j
$\Delta V =$	per unit voltage increment between two successive taps
$\theta =$	shaft angular rotation
$\Delta\theta_s =$	per unit shaft angular increment between two successive taps
$M =$	Mach number
$C_\alpha(m) =$	Non-linear aerodynamic coefficient

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procedure for an arbitrary function can be carried out in a very simple manner, taking into consideration the input impedance of the shunted potentiometer at a predetermined value. The effect of loading, stability of function generator due to change in values of resistors are also discussed. Finally the simulation of many non-linear functions by these function generators in the aerodynamic section of a computer is briefly described.

In mathematical literature[9] several methods are available for approximating non-linear functions by linear interpolation. It is not easy to say which method gives the best approximation and so is the most appropriate one to use. In engineering applications, the method that gives the minimum error of area is preferable since it leads to easy mental visualization. Suppose an arbitrary function is represented by $f(x)$ and its approximation by linear line segments represented by $P_n(x)$, where n denotes the particular segment considered. The condition to be met is

$$\int [f(x) - P_n(x)]\, dx = 0 \quad \ldots\ldots\ldots\ldots (1)$$

The error between the approximation and the arbitrary functions depends upon the solution of differential equations. If one assumes that the error of curve fitting to be the error of the solution, the error may be expressed in either one of the following equations:

$$\xi_a = \frac{\text{Max.}\,|f(x) - P_n(x)|}{\text{Max.}\,|f(x)|} \quad \ldots\ldots\ldots\ldots (2)$$

$$\xi_b = \frac{\int |f(x) - P_n(x)|\, dx}{\int |f(x)|\, dx} \quad \ldots\ldots\ldots\ldots (3)$$

The above measures of error are relative. Equation (2) indicates the degree of approximation for instantaneous values whereas equation (3) gives the accuracy for the entire range. Fig. 1 shows a parabolic function $f(x) = x^2$ approximated by four linear segments. The four ordinates, a, b, c, and d are obtained by the area method formulated by equation (1) for $0 \sim 1$, $1 \sim 2$, $2 \sim 3$ and $3 \sim 4$. From equations (2) and (3), the errors are found to be $\xi_a = 0{\cdot}7$ per cent and $\xi_b = 1{\cdot}9$ per cent.

Biased Diode Function Generators

For detailed discussions, one may refer to a recent paper by Burt and Lange[8]. A unit of somewhat different design is briefly described here in order to show its advantages and disadvantages for ready comparison with the shunted potentiometers in the later part of this article. Basically, a series of diode circuits is used to represent the curve. Each diode circuit represents a segment of the curve. Fig. 2 is a functional diagram of a typical circuit consisting of both positive and negative units. R_1 and VR_1 form a bridge circuit whose output is the difference between e_1 and e_2, i.e. $e_o = k_1(e_1 - e_2)$, where k_1 is a constant. For the positive units, as long as $(R_2/(R_1+R_2))e_{in}$ is less positive than V_1, the diode will not conduct and the output e_o is zero. When $(R_2/(R_1+R_2))e_{in}$ is greater than V_1, the diode conducts and the difference between e_1 and e_2 is proportional to e_{in}: $e_o = k_1(e_1 - e_2) = k_1 k_2 e_{in}$, where k_2 is another constant. k_1 depends on the value of R_f, and k_2 the setting of P_1. If P_1 is precisely centred, $k_2 = 0$. The output remains zero until $(R_2/(R_1+R_2))e_{in}$ exceeds V_1. When $(R_2/(R_1+R_2))e_{in}$ exceeds V_1, the output is proportional to further increases in the input voltage. The slope beyond V_1 is adjustable by means of potentiometer VR_1 and can be made either positive or negative. For negative inputs, negative units are used, the principal is very similar to that for the positive unit.

The setting up of the function generator involves successively determining the break voltages of each of the diodes and adjusting the slope control of the preceding diode each time to give the function value at the break voltage of the succeeding diode.

Function Generation by Shunted Potentiometer

TAPPED POTENTIOMETER

These potentiometers have wires attached to the winding at prescribed points[10]. These wires, called taps, are brought to a terminal strip on the potentiometer case, thus making

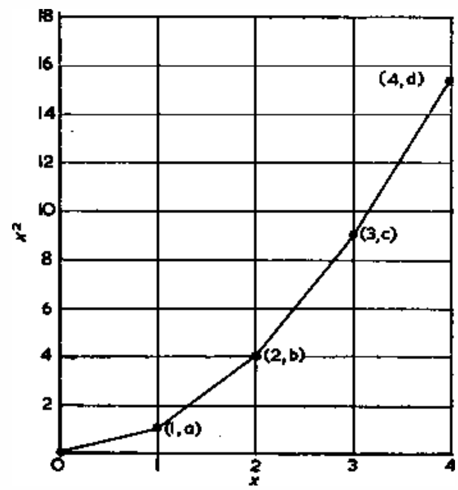

Fig. 1. Parabolic function $f(x) = x^2$ approximated by four line segments

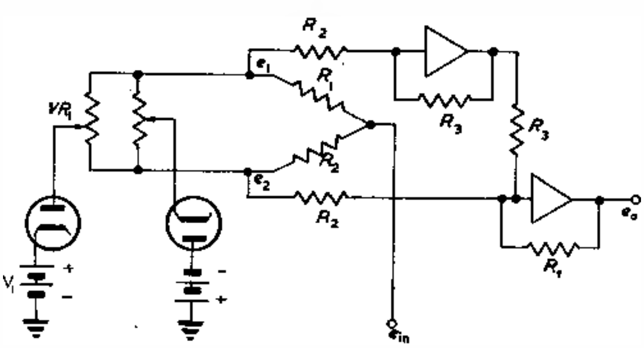

Fig. 2. Basic diode function generator consisting of both positive and negative units

the tapped points externally accessible to electrical connexion from the outside. The potentiometer winding is usually excited by a voltage impressed across the ends of the winding. By attaching resistors of selected value to the various tap terminals, the usual linear voltage variation of the output voltage with output arm position of an ordinary potentiometer is converted into a non-linear variation. However, voltage variation between any two successive taps is still linear, so the desired function is approximated by linear interpolation. It is essential that the output of the shunted potentiometer should feed a load of sufficiently high impedance in order that the desired voltage distribution is not disturbed. In many applications the output arm is controlled by a servo, multiplication of a variable by the non-linear function is accomplished in a manner similar to servo multiplication. Essentially the problem of shunt loading a potentiometer winding in the manner mentioned above reduces to selecting the correct values of the shunt resistors.

MONOTONIC FUNCTIONS

Fig. 3 shows a monotonically increasing function approximated by a sequence of segments. The dependent variable is represented by V since the function is generated by voltage excitation of the potentiometer. The independent variable is represented by θ, which stands for the angular position of the potentiometer output arm. The range of the functional values is restricted to between 0 and 1 after normalization. Both V and θ are expressed in per unit values. Suppose the potentiometer has equally spaced

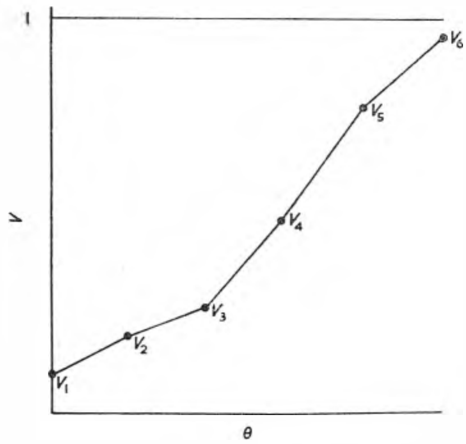

Fig. 3. A monotonically increasing function approximated by a sequence of segments

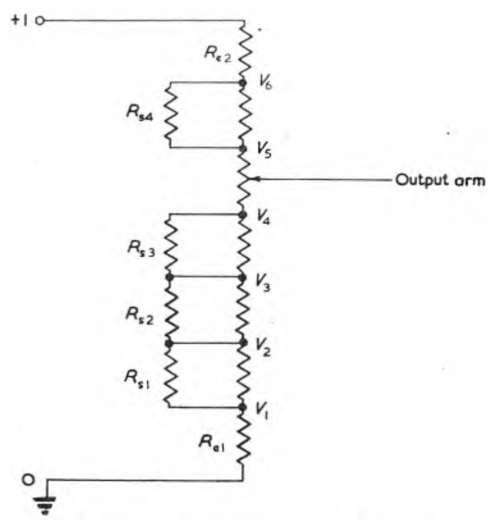

Fig. 4. Circuit representation of Fig. 3

winding resistance values between taps, the resistance unit is then chosen to be the resistance value per winding segment, i.e. R_w/n, where R_w is the total winding resistance of the potentiometer and n is the number of winding sections.

Let the segment that has the largest voltage increment say $V_5 - V_4$ have the total source current I passing through it. Thus $I = V_5 - V_4$ and the input impedance is simply

$$z_i = \frac{1}{V_5 - V_4} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

To obtain the desired voltage distribution at the various taps, two end resistors R_{e1} and R_{e2} are required to provide the necessary low and high end voltages of the potentiometer and the segments must be suitably shunted by R_{s1}, R_{s2} etc. to reduce the current I to appropriate values. This gives rise to the circuit configuration shown in Fig. 4. Thus one has

$$R_{e1} = \frac{V_1}{V_5 - V_4} \qquad R_{e2} = \frac{1 - V_7}{V_5 - V_4}$$

$$R_{s1} = \frac{V_2 - V_1}{(V_5 - V_4) - (V_3 - V_1)} \qquad R_{s2} = \frac{V_3 - V_2}{(V_5 - V_4) - (V_3 - V_2)}$$

$$R_{s3} = \frac{V_4 - V_3}{(V_5 - V_4) - (V_2 - V_1)} \qquad R_{s4} = \infty$$

$$R_{s5} = \frac{V_6 - V_5}{(V_5 - V_4) - (V_6 - V_5)} \qquad R_{s6} = \frac{V_7 - V_6}{(V_5 - V_4) - (V_7 - V_6)}$$

$$\ldots\ldots\ldots\ldots \quad (5)$$

All the above resistances are in per unit values. In actual ohms each one has to be multiplied by (R_w/n).

Arbitrary Function

An arbitrary function shown in Fig. 5 is approximated by 10 sections. The voltage distribution at the taps are indicated and the potentiometer is excited by a $+1$ source.

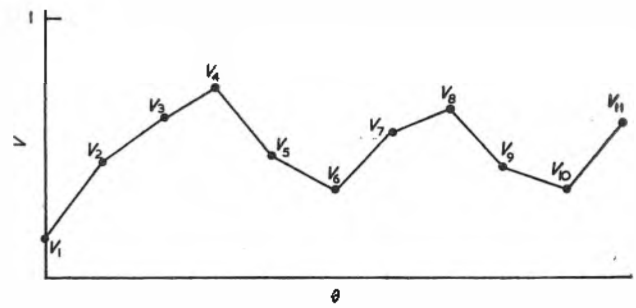

Fig. 5. An arbitrary function approximated by 10 segments

Beside the two end points of the winding, there are two voltage maxima and two voltage minima represented by V_4, V_8, and V_6, V_{10} respectively. To maintain current continuity at these nodes, resistors R_{p1} and R_{p3} are used to feed appropriate currents from the $+1$ source to nodes 4 and 8, while R_{p2}, R_{p4} to draw currents away from nodes 6 and 10 to earth. These four resistors are called 'pull-in' resistors and they are required whenever the function has a change in the sign of the slope. Evidently the function consists of five monotonic sections. In each section there is a segment of largest voltage increment corresponding to the largest current in that monotonic section. These are $V_2 - V_1$, $V_4 - V_5$, $V_7 - V_6$, $V_8 - V_9$ and $V_{11} - V_{10}$. Now by shunting the remaining segments of the potentiometer, the desired voltage distribution of the various taps can be achieved. The corresponding circuit connexion is indicated in Fig. 6. The various resistances in per unit values are determined as follows:

$$R_{e1} = \frac{V_1}{V_2 - V_1} \qquad R_{e2} = \frac{1 - V_{11}}{V_{11} - V_{10}}$$

$$R_{p1} = \frac{1 - V_4}{(V_4 - V_5) + (V_2 - V_1)} \qquad R_{p2} = \frac{V_6}{(V_7 - V_6) + (V_4 - V_5)}$$

$$R_{p3} = \frac{1 - V_8}{(V_8 - V_9) + (V_7 - V_6)} \qquad R_{p4} = \frac{V_{10}}{(V_{11} - V_{10}) + (V_8 - V_9)}$$

$$R_{s1} = \infty \qquad R_{s2} = \frac{V_3 - V_2}{(V_2 - V_1) - (V_3 - V_2)}$$

$$R_{s3} = \frac{V_4 - V_3}{(V_2 - V_1) - (V_4 - V_3)} \qquad R_{s4} = \infty$$

$$R_{s5} = \frac{V_5 - V_6}{(V_4 - V_5) - (V_5 - V_6)} \qquad R_{s6} = \infty$$

$$R_{s7} = \frac{V_8 - V_7}{(V_7 - V_6) - (V_8 - V_7)} \qquad R_{s8} = \infty$$

$$R_{s9} = \frac{V_9 - V_{10}}{(V_8 - V_9) - (V_9 - V_{10})} \qquad R_{s10} = \infty$$

$$\ldots\ldots\ldots\ldots \quad (6)$$

Finally the per unit input impedance of the network to the source is

$$Z_i = \frac{1}{V_{11}-V_{10}+V_8-V_9+V_7-V_6+V_4-V_5+V_2-V_1} \quad \ldots (7)$$

An alternative method of connecting the tapped potentiometer without using shunt resistors is shown in Fig. 7. The desired voltage distribution at various taps is maintained by feeding from the source to nodes 2, 3, 4, 7, 8, 11 and drawing currents from nodes 5, 6, 9, 10 to earth. The resistance values are easily seen to be

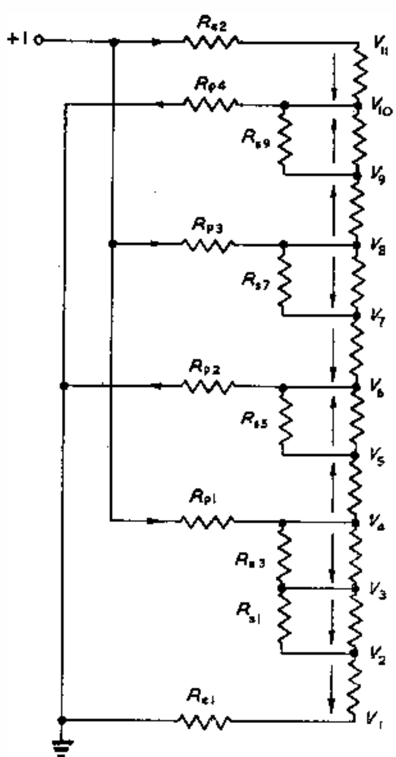

Fig. 6. Circuit representation of Fig. 5

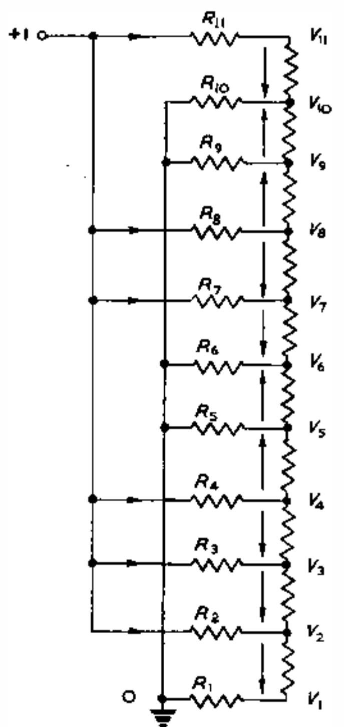

Fig. 7. An alternative circuit representation of Fig. 5

$$R_1 = \frac{V_1}{V_2 - V_1} \qquad R_2 = \frac{1-V_2}{(V_2-V_1)-(V_3-V_2)}$$

$$R_3 = \frac{1 - V_3}{(V_3-V_2)-(V_4-V_3)} \qquad R_4 = \frac{1 - V_4}{(V_4-V_3)+(V_4-V_5)}$$

$$R_5 = \frac{V_5}{(V_4-V_5)-(V_5-V_6)} \qquad R_6 = \frac{V_6}{(V_5-V_6)+(V_7-V_6)}$$

$$R_7 = \frac{1 - V_7}{(V_7-V_6)-(V_8-V_7)} \qquad R_8 = \frac{1-V_8}{(V_8-V_7)+(V_8-V_9)}$$

$$R_9 = \frac{V_9}{(V_8-V_9)-(V_9-V_{10})} \qquad R_{10} = \frac{V_{10}}{(V_9-V_{10})+(V_{11}-V_{10})}$$

$$R_{11} = \frac{1 - V_{11}}{V_{11} - V_{10}} \quad \ldots\ldots\ldots\ldots (8)$$

The input impedance of the network is the same as in the preceding connexion.

Design Based on Network Tearing

In the design of shunting circuits for tapped potentiometers, several factors have to be taken into consideration. Since tapped potentiometers are supplied with their voltage from the output of an amplifier, the input impedance of the shunted tapped potentiometer must be suitably chosen beforehand. From equation (7) it may happen, particularly for functions that are nonmonotonic, that the input impedance falls below a desired value. If so, a padding resistor is required to be inserted in series with the driving source. This will reduce the scale, which must be restored by a corresponding compensating gain in another place in the analogue. The output of the potentiometer usually feeds a 1MΩ input to an amplifier. If the function being represented only covers a restricted part of the range 0 to 1 say from 0·9 to 1, the entire tapped potentiometer may have to be raised from earth by a large value resistor, then the effect of loading by 1MΩ would be appreciable. Usually the function has to be transferred to the entire 0 to 1 range by adding and subtracting a constant and renormalizing. This increases the number of amplifiers required. If the tapped potentiometer is to be custom-designed for the particular non-linear application, it is usually necessary to space the taps closely in the region of maximum function curvature. The analysis in previous section based on winding segments of equal resistance values will no longer hold.

A powerful method in the analysis and synthesis of networks is the method of tearing. This idea originated by Kron has many interesting applications. It will be seen that by tearing the complex network of the shunted potentiometer into primitive circuits, the design of the tapped potentiometer function generator which will have a final input resistance of a predetermined value becomes extremely simple and straightforward.

Take the non-linear function of Fig. 5 and replot it in Fig. 8 with per unit voltage increments between two successive nodes and the corresponding per unit shaft angle increments labelled. If $\Delta\theta_{s1} = \Delta\theta_{s2} = \Delta\theta_{s3} = \ldots$, one has the special case of equally spaced taps. A useful graphical representation that leads naturally to the idea of tearing is shown in Fig. 9. It is a resistance diagram corresponding to the function diagram. In the graphical representation the potentiometer resistance element follows exactly the straight line approximation of the non-linear function. Hence the five monotonic sections can be electrically separated, each pull-in resistor is replaced by two resistors having twice its original value shown in Fig. 10. Each branch is now excited by the same source voltage. Since the shunting resistances are yet to be determined, a degree of freedom exists to permit one imposing a condition such that the branch currents are all equal. In other words, all branches will have equal resistance values, say R_b, and the input impedance is simply R_b/b, where b is the number of branches or the number of monotonic sections. This condition greatly facilitates the calculation of shunting resistances, because the total resistance in ohms of each segment in a branch is equal to the per unit voltage increment of that segment multiplied by the branch resistance, i.e., $R_b = bR_i$.

The end resistors can be computed from

$$R_e = \Delta V_e b R_i \quad \ldots\ldots\ldots\ldots (9)$$

and the pull-in resistors from

$$R_p = \frac{\Delta V_p b R_i}{2} \quad \ldots\ldots\ldots\ldots (10)$$

The total resistance of each segment of the potentiometer is made up of the winding resistance $\Delta\theta_s R_w$ and the shunt resistor R_s, hence

$$1/R_s = \frac{1}{\Delta V_s R_i} - \frac{1}{\Delta\theta_s R_w} \quad \ldots\ldots\ldots (11)$$

For a potentiometer with equally spaced taps, $\Delta\theta_s R_w$ becomes R_w/n and b equals to 1. After normalization

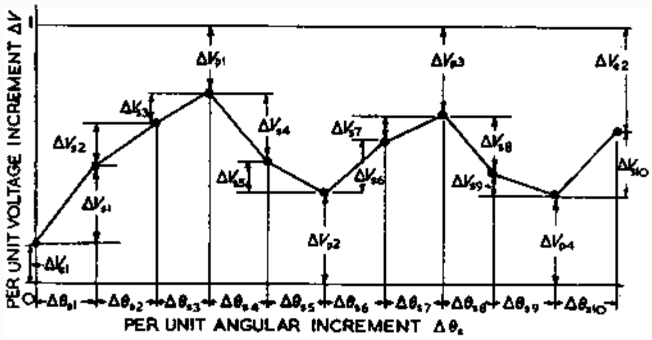

Fig. 8. A replot of Fig. 5 showing the per unit voltage increments between two successive nodes and the corresponding per unit shaft angle increments

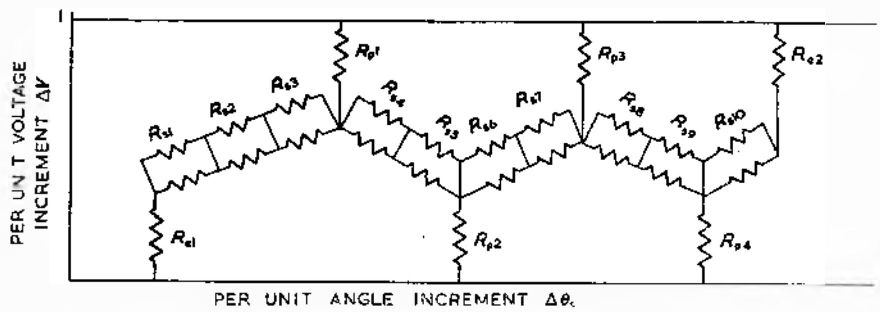

Fig. 9. Resistance diagram corresponding to Fig. 8

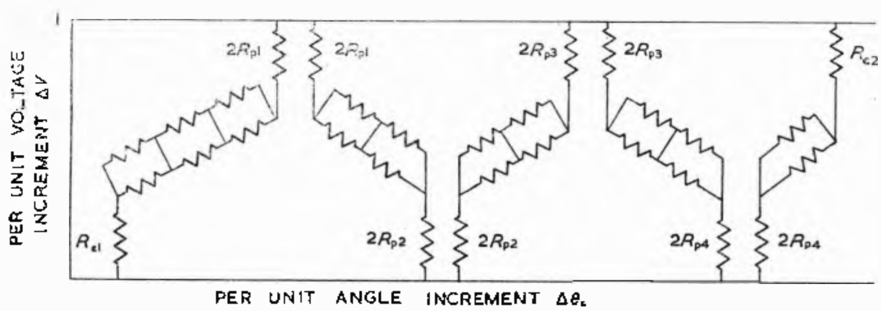

Fig. 10. Separation of Fig. 8 into five monotonic sections

equation (11) is reduced to

$$1/R_s = \frac{1}{\Delta V_s R_i'} - 1 \quad \ldots\ldots\ldots\ldots (12)$$

Where R' is equal to $R_i/[(R_w)/n]$, the correlation of equation (12) with equation (5) is obvious.

For a given input impedance, the winding resistance of the potentiometer to be shunted must be greater than a minimum value due to the fact that shunting circuits can only reduce the net resistance and the potentiometer resistance must be chosen high enough to allow for this reduction. This also depends upon the value of the maximum slope of the straight line segment. Thus rewriting equation (1)

$$1/R_s = \frac{\Delta\theta_s R_w - \Delta V_s b R_i}{\Delta V_s b R_i \,\Delta\theta_s R_w} \quad \ldots\ldots\ldots (13)$$

Since R_s cannot be negative, one has

$$R_w \geqslant (\Delta V_s/\Delta\theta_s)\, bR_i \quad \ldots\ldots\ldots\ldots (14)$$

In actual design, if the input impedance has a predetermined value, one should determine the value of the slope of the straight line segment having the maximum slope and then select a tapped potentiometer having a winding resistance greater than the value calculated from equation (14).

Effect of Loading

Consider a potentiometer representing a monotonic function as in Fig. 3 with the maximum voltage normalized to 1. A resistive load R_L is corrected across the j^{th} tap and earth. Let V_j be the voltage at the j^{th} tap when unloaded and V_j' the voltage at the same tap when loaded by R_L, then

$$\Delta V_j = V_j - V_j' = V_j'(Z_j/R_L) \quad \ldots\ldots\ldots\ldots (15)$$

Where Z_j is the output impedance at the j^{th} tap, Z_j is composed of two impedances in parallel, one between j^{th} tap and positive source, the other between j^{th} tap and earth. In terms of V_i and the input impedance Z_i, one has

$$Z_j = (1 - V_j)\,V_j Z_i \quad \ldots\ldots\ldots\ldots\ldots (16)$$

Since

$$R_w > Z_i(1 - V_i) \quad \ldots\ldots\ldots\ldots\ldots (17)$$

Also $V_j' > V_j$, then

$$\Delta V_j < \frac{(1-V_j)}{(1-V_j)} V_j^2 [(R_w)/R_L] \quad \ldots (18)$$

The maximum value of $(1-V_j)V_j^2$ is 4/27 which occurs when $V_j = (2/3)$, thus, if R_i is 50 times greater than R_w, the reduction in functional value at j^{th} tap due to loading is always less than 1 per cent should the function extend over a range from unity down further than 0·7. The same conclusion would apply to nonmonotonic functions. For a 20kΩ tapped potentiometer loaded by a 1MΩ input of an amplifier, the loading effects will all be less than 1 per cent, except when the functional range on one monotonic section is very small.

Influence of Resistor Tolerance on the Stability of Function Generator

It is interesting to study how the tap voltages will change, if the resistors of the network change value. Using resistor units of ±1 per cent manufacturing tolerance, the errors are very small values indeed. Usually an accuracy in the range of 1 part in 1 000 can be expected with this type of network depending upon the scale and geometry of the set up. Assume the deviation $\partial R = \rho R$ of the actual resistance values from their nominal values R are randomly distributed and that the distribution is given by a normal frequency curve with a standard deviation $\sigma = \sigma(\rho)$, where $\rho = (\partial R/R)$. Now let one resistor between any two successive taps say, P_1 and P_2, be changed by ∂R to $R + \partial R = R(1+\rho)$, the voltage difference between P_1 and P_2 increases to $\Delta V + \partial V$, causing a change in the potential at the point P_1. The voltage differences betwen P_1 and other nodes conductively connected to P_1 remain the same as before, except for a negligibly small correction term which may be omitted. Summing up the branch currents flowing into P_1 to zero, one may obtain an expression for ∂V in terms of ρ and other known quantities. This is the local error due to the change of one resistance. As the errors ρ were assumed to be statistically distributed, a normal error distribution for ∂V can be expected with a standard deviation $\sigma(\partial V)$, obtained by replacing ρ by $\partial(\rho)$ and ∂V by $\sigma(\partial V)$ in the local error expression. To obtain the local error of the potential at the point P_1 due to the presence of tolerance errors of all resistances joined in the point P_1, the voltage changes due to each of the resistances have to be superimposed.

Next, one has to consider the effect of the statistical accumulation of the local errors in the whole circuit, leading to an average error of the voltage distribution which is superimposed on the local error. This error-reducing statistical property possessed by the resistance network makes it a useful and attractive research and design tool. For the

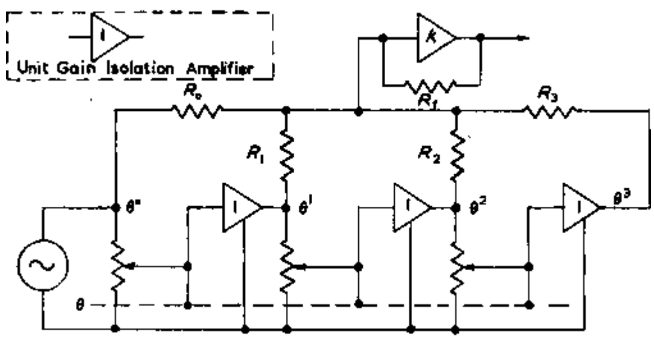

Fig. 11. Power series representation by cascading ganged linear potentiometers

monotonic function in Fig. 4, it is obvious that the voltages at the output arm will have its maximum possible increase if all resistances below it increase and above it decrease. Calling the upper total resistance R_1 and the total lower resistance R_2, the per unit increase in voltage at the output arm is

$$dV = [(dR_2/R_2) - (dR_1/R_1)]\left[\frac{R_1R_2}{(R_1 + R_2)^2}\right] \ldots\ldots (19)$$

Since the maximum value of $R_1R_2/(R_1+R_2)^2$ is $1/4$, $dV \leqslant 0{\cdot}02 \times 1/4 = {\cdot}005$, which is $\frac{1}{2}$ per cent of the full excitation voltage. Clearly this is the most pessimistic result not obtainable in practice.

Application to Aerodynamic Problems

It may be mentioned that to simulate the aerodynamic force and torque equations by means of tapped potentiometers in an analogue computing machine, many non-linear functions must be generated in the aerodynamic section of the computer. The non-linear function terms are usually of the form $\alpha C_{\alpha(M)} + \delta C_{\delta(M)} + \ldots\ldots$ where $C_{(M)}$ is presented in graphical form and is a function of Mach number M.

The aerodynamic coefficients are simulated by suitably shunted potentiometers where output arms are controlled by Mach servos. The input to the tapped potentiometer is supplied by an amplifier generating a variable voltage, say α or δ. Essentially the servo-driven tapped potentiometers is a modified servo multiplier followed by summing networks. Accuracy achieved by this method is in the order of 1 or 2 per cent.

Non-linear aerodynamic functions can also be generated by means of cascaded ganged linear potentiometers. The degree of precision obtained depends on how well the non-linear function is approximated by a power series expansion. Fig. 11 shows three linear potentiometers connected to provide a series involving powers of shaft rotation θ.

$$V(\theta) = C_0 + C_1\theta + C_2\theta^2 + C_3\theta^3 \ldots\ldots\ldots (20)$$

The voltage source gives the constant term, while the outputs of successive isolation amplifiers give the successive powers of θ. The scale factor C of each of these terms is established by the combining resistor which goes to the voltage summation point. The summation of terms is achieved by the feedback amplifiers. From practical instrumentation considerations, this method is not as good as tapped potentiometers in solving aerodynamic problems because to obtain the same order of accuracy, the polynomial method would require considerably more electronic equipment.

Conclusions

The shunt loading technique described in this article appears to be the most suitable means of generating non-linear functions in analogue computers for the simulation of aerodynamic equations. It has distinct advantages over diode function generators and polynomial generators in flexibility, simplicity, reliability and repeatability. Using high precision wire-wound resistors, the tapped potentiometers do not require a calibration procedure for checking the slopes and break point voltage as in diode function generators. The calculations of the shunting circuits are simple and the speed of operation is favourably compared with that of a diode function generator.

For special purposes, shunting techniques can be combined with other techniques. For example, the non-linear terms achieved by shunting technique can be added to the power series expansion by cascaded ganged potentiometers through appropriate mixing resistors. Smoothly varying non-linear functions such as tangent, secant, square root, square and reciprocal, etc., can be reproduced quite accurately by attaching a selected shunt resistor from the output arm to the appropriate end of the potentiometer winding. This entails proper selection of the ratio of the shunt resistor to the winding resistor, and by use of only a selected portion of one of these curves. Furthermore, if the tapped potentiometers are not highly linear, voltage clamping of tap points can also be used with advantage to represent non-linear function with steep slope.

Acknowledgments

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A Vacuum Plant for the Frequency Calibration of Quartz Vibrators

By J. Green*, L. Holland* and B. D. Power*

A vacuum evaporation plant is described for the final frequency calibration of quartz vibrators. A single crystal is mounted in a small chamber exhausted by an air cooled diffusion pump charged with silicone oil. High pumping speed and rapid pumping times are obtained by connecting the diffusion pump directly to the chamber and dispensing with the high vacuum valve commonly used for isolating the diffusion pump from atmosphere. The hot diffusion pump fluid is exposed to the atmosphere at the end of each cycle and oil losses during pre-pumping by the rotary pump prevented by a special baffle in the backing line. A method has been devised to prevent silicone oil mist from entering the chamber when the system is open to atmosphere. A crystal can be brought to frequency with gold or aluminium electrodes in from one to two minutes. The apparatus can be automatically controlled by a process timer.

HIGH frequency crystal vibrators of the kind shown in Fig. 1 are brought to the desired frequency by controlling the mass of the electrodes deposited on their surfaces. It is common practice to deposit by cathodic sputtering or vacuum evaporation the metal electrodes on a number of crystals simultaneously, known as base coating, and then to bring each crystal to the correct frequency by evaporating a small amount of metal on one of the electrode surfaces while measuring the crystal frequency or its difference from a known standard.

Occasionally crystals are brought directly to frequency by dispensing with the base coating operation. Although the coating process is simplified it is more difficult to provide a well defined edge to the evaporated metal electrode when it is deposited in a single operation. This is because the mask inserted in the vapour beam for defining the shape of electrode must not rest on the surface of the oscillating crystal, and a graduated shadow of the edge of the mask may be produced in the deposit. Such an effect depends on the emitting area of the vapour source and the distances between the source, mask and crystal. Also, if either the gas pressure or evaporation rate is high, atomic collisions may occur in the vapour beam causing the path of the vapour atoms to bend past the edge of the mask.

Crystals have been brought to frequency in a single operation using a small crystal to mask distance in the plant described below. However, another problem experienced in using this method was that the vapour source required frequent recharging, because of the greater mass of metal which required evaporation when the whole electrode was deposited in a single operation. Thus, most manufacturers find it preferable to base coat using a mask in contact with the crystal surface, followed by frequency adjustment using a mask to confine the added metal within the base coated area.

The methods adopted for metallizing crystal vibrators have been discussed by Holland[1]. The prime purpose of this article is to describe a simple evaporation plant which has been developed for the rapid frequency calibration of crystal vibrators either in a single operation or after base coating. The pumping system used in the calibration unit to be described possesses several novel features. Its use is not intended however to be limited to the coating of crystal vibrators and it has been recently employed for exhausting cathode-ray tubes during the aluminizing operation.

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Pumping System

Design Problems

There are several problems to be solved in designing a pumping system for use in the rapid frequency adjustment of quartz vibrators and these are discussed below. The pumping system generally used for vacuum evaporation consists of an oil diffusion pump backed by a rotary pump.

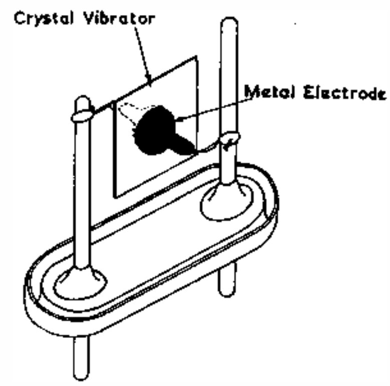

Fig. 1. Typical construction of quartz vibrators for high frequency use

A high vacuum valve is mounted between the diffusion pump and chamber and an isolation valve in the diffusion pump backing line. A valved by-pass pipe line is connected between the rotary pump and chamber to permit rough pumping of the chamber before the backing and high vacuum valves are opened for fine pumping. This valve system has come into general use because it allows the diffusion pump to be kept at working temperature and under vacuum when the chamber is opened to atmosphere. It is essential that the hydro-carbon fluids used in diffusion pumps should not be exposed to the atmosphere while hot, and although this is possible with the more chemically stable silicone oils, there are practical difficulties in preventing the oil from being carried into the rotary pump during the initial stages of pumping, as discussed below.

It has been current practice to metallize quartz vibrators in plant fitted with the conventional valve system described above, and several different methods of controlling the valves have been used for obtaining rapid processing times. Thus in one plant the valves are magnetically operated and in another three chambers are rotated on a sliding seal to pass through the loading, roughing and fine pumping stages, see for example Awender *et al.*[2] Holland[1] has used a standard evaporation plant for frequency calibration by mounting a perspex turret head carrying a number of crystals inside a glass bell jar, each crystal was rotated over a vapour source for frequency adjustment. More recently the turret heads have been made of polystyrene because it has negligible moisture content which reduces the degassing. The apparatus has proved reasonably satisfactory, but suffers from the defects that the mechanism in the chamber is complex and it is not possible to directly connect the crystal to an external oscillator without using long leads. An oscillator can be mounted in the evaporation chamber but this imposes an extra pumping load on the diffusion pump.

Attention was therefore given to the construction of a small vacuum coating plant in which crystals could be brought to frequency one at a time without either a rotary mechanism or long crystal connexions. It was found however that any attempt to use a high vacuum valve with a small capacity chamber seriously lengthened the exhaustion time to a pressure of $0{\cdot}1\mu$ Hg, which is required for the evaporation of pure aluminium films. This was because the tubulation of the valve lowered the pumping speed available at the chamber and also the added surface area of the valve increased the rate of degassing. Thus a metal chamber of 250cm^3 capacity could not be exhausted to $0{\cdot}1\mu$ Hg in less than five minutes using an oil diffusion pump with a speed of 75 l/sec and a combined high vacuum valve and baffle giving a baffled speed of 50 l/sec. The volume and surface area of the high vacuum valve used with the 2¼in pump was some four times greater than that of the evaporation chamber. If the valve was removed and an oil vapour baffle inserted in the pump so that the final speed was 35 l/sec the chamber could be exhausted to the same pressure in under one minute, i.e. providing the diffusion pump was at working temperature. In both cases the rotary pump used had a displacement of 50 l/min.

To obtain rapid pumping times the diffusion pump must be kept near to or at working temperature, although a high vacuum valve cannot be used, and before such a valveless system can be employed the following problems must be solved:

(1) Silicone fluid is removed from the diffusion pump into the rotary pump in the form of an oil mist when air at high pressure is exhausted through a hot diffusion pump.

(2) The heater on a diffusion pump must be switched off when the pump is open to atmosphere for long periods otherwise the oil reaches a high temperature and boils uncontrollably when the gas pressure is reduced. This results in a cloud of oil vapour entering and contaminating the vacuum chamber.

(3) Traces of oil mist pass through the oil baffle and contaminate the vacuum chamber when the hot diffusion pump is exposed to the atmosphere.

(4) It is customary to water-cool the high vacuum valve on conventional pumping systems so that it acts as a cold sink for oil vapour. Without a cooled zone the oil vapour molecules entering the chamber during pumping may condense on the chamber fittings.

The first three problems can be overcome by fitting a water cooling coil on the pump boiler so that the oil can be quickly cooled before air is admitted to the chamber. This technique is still widely used particularly in America, but rapid pumping cycles can only be obtained by using exceptionally high rated heaters on the diffusion pump to obtain a short warming up period, after which part of the heater input may be switched off. It is doubtful, however, whether pumping periods of about one minute from atmosphere to $0{\cdot}1\mu$ Hg could be obtained with this method; particularly since a period for water-cooling must be allowed for after the heaters have been disconnected before air can be admitted. The specially designed pumping system used in overcoming the foregoing difficulties is shown in Fig. 2 and described below.

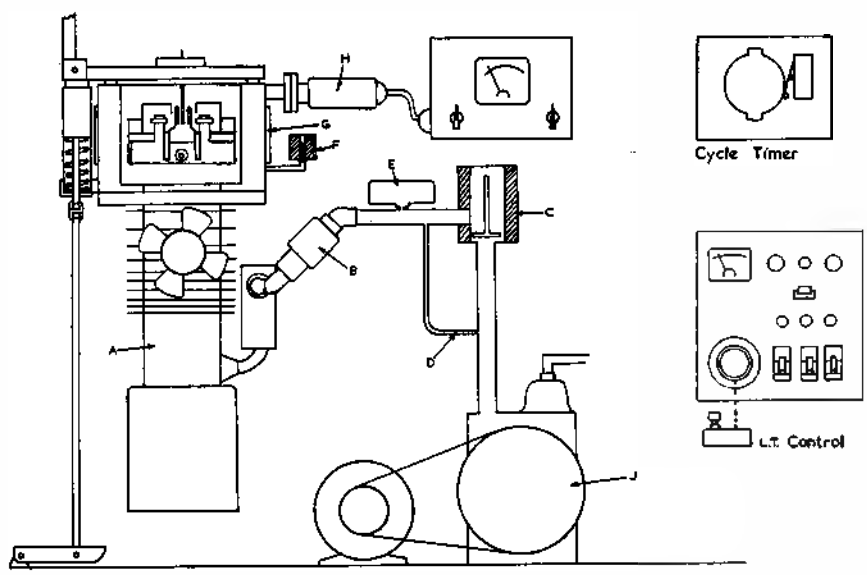

Fig. 2. Pumping system and calibration chamber arrangement

A—air cooled diffusion pump, B—fluid economizer, C—magnetic isolation valve, D—bleed line, E—vacuum pressure switch, F—air admittance valve, G—chamber heater, H—Philips gauge head, J—rotary pump

Fluid Economizer

Oil losses arising from item (1) due to gas at a high pressure flowing through the pump, can be greatly reduced with a special mist baffle developed by Barrett and Power[8] for use in the backing line of a diffusion pump. If the mist laden air stream is divided into a number of small streams which are accelerated to a suitable velocity and directed against a surface so that each stream must turn sharply, then the mist droplets tend to be separated centrifugally during the sharp change of direction and precipitated on the target. Also, the turbulence induced in each of the air streams during acceleration tends to make the mist droplets coagulate before the change of direction occurs and thus improve the efficiency of centrifugal separation. A baffle designed on this principle can easily be made from two plates which have been drilled to provide a number of apertures. The pair of plates are placed in the air stream with a small gap between them. The leading plate is drilled with small holes to provide a restriction and thereby produce the desired acceleration of the gas. The apertures in the down-stream plate are of large size and of a combined area such as to allow gas to pass freely without further acceleration. The apertures in the two plates are staggered so that the gas passing through the first plate impinges on the wall of the second. The gas stream bends sharply and the oil droplets are precipitated on the plate from which they drain and flow back into the diffusion pump boiler. The size of the apertures in the front plate will depend on the rate of flow of gas through the pump. This being determined by the displacement of the rotary pump and resistance of the backing line.

The economizer fitted to the calibration plant consists of two sets of drilled plates followed by a pair of wire gauze disks which collect any large droplets of oil remaining in the air stream or any liquid which may be blown off as a spray from the baffle plates.

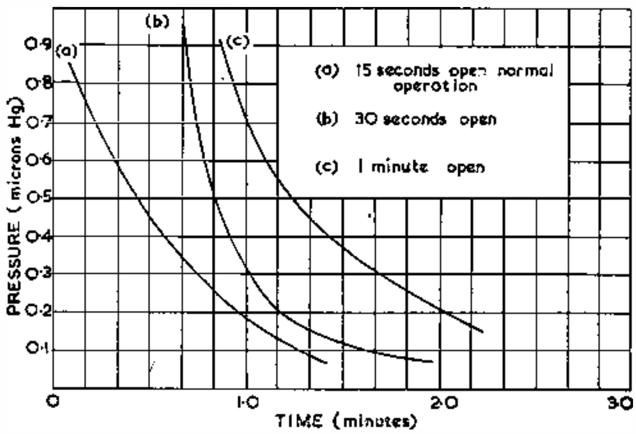

Fig. 3. Pumping time-pressure curves

The pumping time to a given pressure depends upon the period for which the chamber is open to atmosphere, because the pump heaters are switched off and the pump oil can cool to different temperatures. The time-pressure curves have been taken with the system continuously cycling and with exposure period of 15, 30 and 60sec

With the pumping combination used (2¼in air cooled diffusion pump and 50 l/min rotary pump) the silicone oil lost when exhausting a 250cm³ chamber is 15cm³ per 1 000 pumping cycles; that is with the diffusion pump open to atmosphere for a period of 15sec between cycles. The diffusion pump is charged with 75cm³ of oil and will continue to pump until the charge has fallen to some 25cm³. Thus the system can be operated at full production for nearly a fortnight before being recharged with oil.

Superheating

To prevent the diffusion pump fluid from superheating when exposed to atmosphere, as described in item (2) above, the boiler heaters must be switched off. However, this allows the boiler to cool thus slowing up the subsequent pumping cycle. An air cooled diffusion pump is used in the plant described, and the cooling fan is switched off together with the heaters when the air admittance valve is opened. By this means it is possible to reduce loss of heat from the pump boiler. However, some cooling does occur and the exhaustion time to a given pressure is dependent on the period for which the heaters are off, i.e. the chamber is open to atmosphere, as shown in Fig. 3. In practice the chamber will only be open to the atmosphere for short periods of 15 to 30sec, and thus there is no great risk of either superheating or undue loss in working temperature, whether the heaters are left on or switched off respectively. However, an operator may inadvertently leave the pumping system switched on and open to atmosphere so that the oil could superheat and it is better to disconnect the heaters at the end of a pumping cycle. For safety purposes the heaters are controlled in the calibration plant by a simple pressure operated switch which switches the power on when the pressure has fallen to about 20mm Hg as happens within the first few seconds of rough pumping. Superheating may still of course occur at this pressure. However, one need only guard against the operator switching on the diffusion pump with the chamber open to atmosphere. The vacuum seals are reliable and unlikely to develop leaks and if the chamber is correctly closed the gas pressure falls well below 20mm Hg in a few seconds. Obviously if the pressure does not sufficiently fall to operate the safety switch then a period will elapse after correcting the fault before the pump boiler again reaches its working temperature.

Bleed Line

To keep the diffusion pump oil losses to a minimum the rotary pump should be switched off when air is admitted to the vacuum chamber. The fluid economizer then only functions during the rough pumping on the subsequent cycle. However, if the rotary pump is switched off when the system is opened to atmosphere a little oil mist may rise and enter the chamber as described in item (3) above. It was found that a very small flow of air through the diffusion pump was sufficient to prevent oil mist from entering the chamber. Such a limited gas flow was obtained by fitting a magnetic isolation valve on the rotary pump and bypassing the valve with a bleed line. When the chamber is at atmospheric pressure the rotary pump is left in operation with the magnetic valve closed so that a small quantity of air flows through the bleed line[4].

The air flow through the bleed line must not be too great otherwise the air admittance valve will not be able to raise the work chamber to atmospheric pressure. The air flow required to suppress oil mist in a 2¼in diffusion pump is extremely small being about 2 l/min, and this can be obtained using a bleed line made from a one foot length of 1/16in i.d. copper tubing.

Oil Back-Streaming

The diffusion pump is fitted with an optical oil baffle in the pump mouth to reduce back-streaming of oil molecules at reduced pressures. However, the oil molecules entering the chamber may condense on the surrounding walls since the temperature of the air cooled pump and chamber are of similar order and thus the vapour pressure of the oil molecules in the chamber may reach the saturated value. By slightly raising the chamber temperature above that of the pump baffle the oil molecules entering the chamber can be prevented from condensing. Further, if the chamber temperature is slightly above ambient the amount of water vapour condensed on its surface is reduced when the chamber is exposed to atmosphere. The chamber temperature is raised a few degrees by winding a few turns of resistance wire on its outside and connecting the element in series with the diffusion pump heaters.

Pressure Measurement

The accurate measurement of the gas pressure in a vacuum chamber depends upon the speed of the tubulation connecting the gauge head to the chamber, and the position chosen for connecting the gauge in relation to the pumping aperture. The restriction due to the gauge tubulation becomes very important when the chamber is of very small capacity, because the gas pressure may fall more rapidly to the ultimate value in the chamber than in the vacuum gauge; particularly if this is of the Philips' type which tends to release gas when first switched on.

The ideal position for the gauge head would be inside the vacuum chamber, but the electrodes would need to be shielded from the vapour source and this introduces other difficulties. However, fairly accurate indication of pressure was obtained on the plant under discussion by mount-

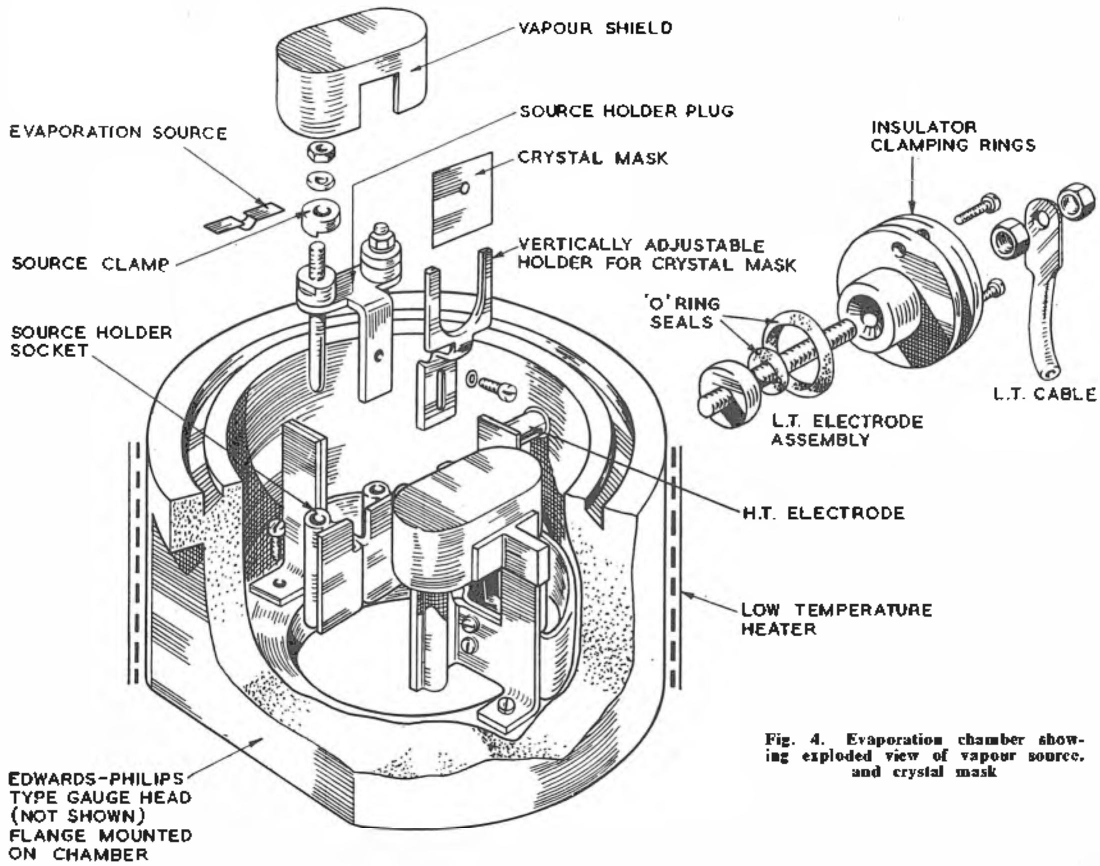

Fig. 4. Evaporation chamber showing exploded view of vapour source, and crystal mask

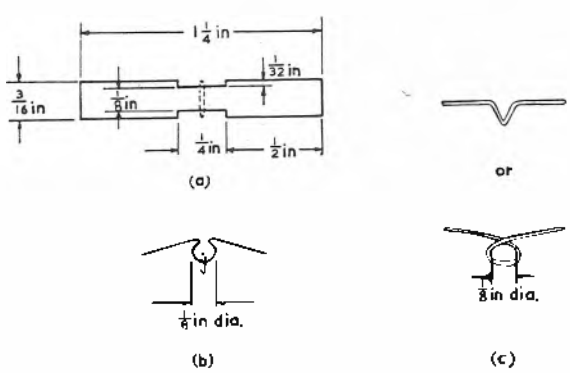

Fig. 5. Evaporation sources

ing the vacuum gauge on the wall of the vacuum chamber so that the tubulation was less than ½in in length and 1in in bore.

Pumping Performance

The speed with which a vacuum chamber can be exhausted depends with the valveless system, on the period for which the diffusion pump is open to atmosphere. The air release valve has a 1/16in diameter opening and is magnetically controlled. It takes about 7sec to raise the chamber to atmospheric pressure and about 8sec for removal and reloading of the crystals. With 15sec exposure periods the chamber can be exhausted to $0{\cdot}5\mu$ Hg in 30sec and $0{\cdot}1\mu$ in about 1min, so that a crystal could be processed in from one to two minutes. The pumping times for different exposure periods, viz., 15, 30 and 60sec, are shown in Fig. 3. The gas pressure chosen for evaporation depends upon the nature of the metal being deposited. Aluminium should be evaporated at pressures less than $0{\cdot}5\mu$ Hg, preferably $0{\cdot}1\mu$ Hg, because of the readiness with which it reacts with residual oxygen and water vapour to form an oxide. Gold is not so reactive and for finishing to frequency may safely be evaporated at pressures of the order of 1μ Hg.

However, if gold is evaporated in a single operation lower pressures are essential, because when gold is evaporated from tungsten or molybdenum heaters at a high gas pressure, the deposit is discoloured and contaminated with tungsten or molybdenum oxides respectively.

Evaporation Facilities

Evaporation Chamber

An exploded view of the evaporation chamber is shown in Fig. 4. It consists of a small metal cylinder, grooved at both ends to take 'O' ring vacuum seals. The inside of the chamber is polished and bright chromium plated. The air-cooled diffusion pump is mounted directly under the chamber. The chamber is closed at the top with one of two lids which are mounted at right-angles to each other and pivoted at the centre. The lids can be rotated on the central rod. Each lid carries a crystal two pin holder made from polystyrene with lead in electrodes for connexion to an external oscillator. The lids are spring loaded so that when air is admitted to the chamber they rise free from the chamber. At the end of a cycle the lids are rotated so that the lid carrying a new crystal is moved over the chamber and then pulled into position by a pedal as shown in Fig. 6. This arrangement permits the operator to load a crystal while another is undergoing exhaustion. By mounting the lids at right-angles the unused lid is in the most convenient position for unloading and loading. The crystals are handled with a pair of tweezers designed to clamp on the base of the holders. Two insulated lead in electrodes can be fitted into the chamber for operating a glow discharge, which is often useful for pre-cleaning purposes.

VAPOUR SOURCES

Two vapour sources are mounted on plug-in electrodes inside the chamber, so that crystals can be coated on both or one face as desired. The source heaters shown in Fig. 5 are made either from tungsten wire for evaporating aluminium or molybdenum foil for evaporating gold. The filament source is loaded with a loop of aluminium wire at the centre which melts to form a small emitting surface. The vapour masks are made from ·004in thick molybdenum foil. Each of the vapour sources is covered with stainless steel shields which prevent evaporated coatings from being deposited on the chamber and which can be chemically cleaned.

CONTROL OF THE EVAPORATION RATE

Careful control of the evaporation rate is required for accurate adjustment to frequency. If the vapour source is raised quickly from room temperature to the evaporation point (≃1 200°C) during the calibration period then the evaporation rate may rise excessively and the required frequency be exceeded. To avoid this the vapour source is raised to dull red heat immediately the pumping cycle commences. The source then only needs to be raised by some 400°C to reach the evaporation temperature permitting greater control of the evaporation rate.

The source heater circuit consists of a 'Variac' transformer controlling a low tension transformer supplying current to the evaporation heater. A single pole two way microswitch is connected in the input of the l.t. transformer. When the microswitch is depressed the 'Variac' is disconnected and full mains voltage applied to the l.t. transformer. The pre-heat temperature of the source is adjusted by the 'Variac.' In practice a crystal is brought to frequency by pulsing the source using the microswitch.

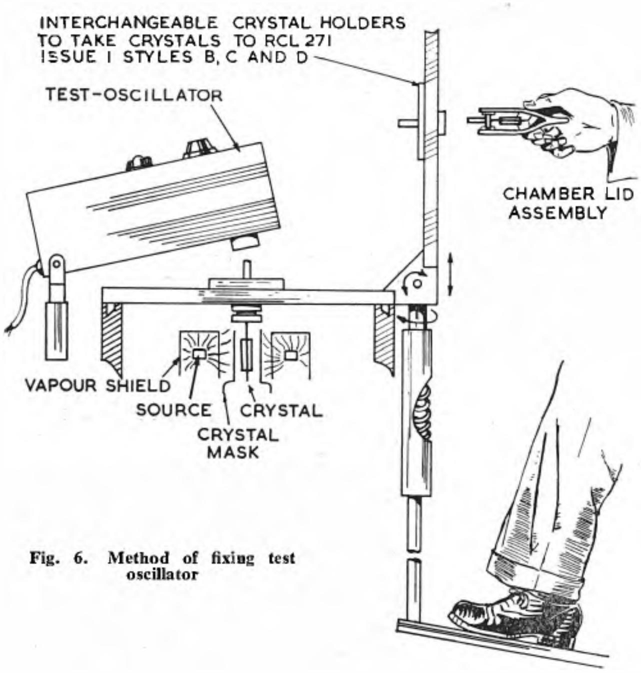

Fig. 6. Method of fixing test oscillator

OSCILLATOR CONNEXIONS

The crystal holder consists of two stainless steel sockets embedded in a polystyrene block with an 'O' ring vacuum seal. Connexion to an external oscillator is made by the protruding holder pins. The total length of connexion from the crystal to the oscillator is about 1in.

The test oscillator is mounted in a box which swivels at the back of the vacuum chamber and which is made to rest on the lead-in electrodes during evaporation when the lid is closed (see Fig. 6).

Automatic Control

The pumping system can be made to cycle continuously by means of a two way switch operated by an electric clock. At the commencement of the pumping cycle the switch applies power to: the magnetic valve on the rotary pump, the diffusion pump heater and cooling fan (i.e. providing the pressure safety switch has operated); at the end of 45sec the power is removed from the foregoing and the magnetic air release valve is opened for 15sec. The rotary pump is switched on at the commencement of the work and operates continuously. The evaporation sources can also be operated automatically, but this would not seem worthwhile unless the evaporation was terminated by a relay operated from the test oscillator. In that case the metal deposition would probably need to be cut off by a magnetically controlled shutter, because it would not be easy to pulse the source as with manual control, and the thermal inertia of the source could allow the evaporation to continue after the heater current was switched off. The vapour sources have been automatically fired by a pressure controller on the Philips gauge but if the atmospheric exposure periods are kept constant then a time control is adequate, because the time-pressure relation is consistent.

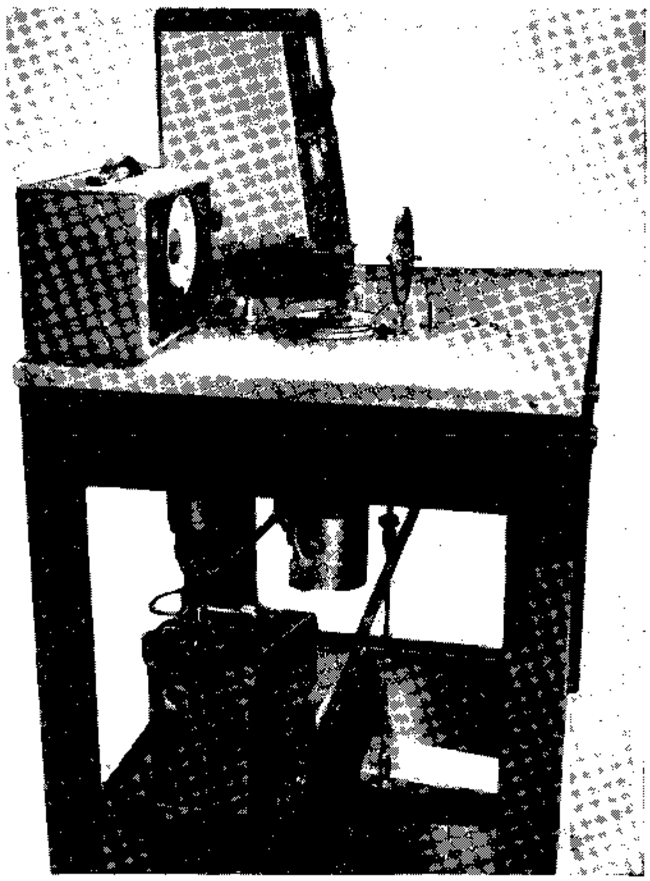

Fig. 7. General layout of frequency calibration plant

A photograph of the completed plant is shown in Fig. 7.

Acknowledgments

Acknowledgments are made to Mr. A. S. D. Barrett, Technical Director of Edwards High Vacuum Ltd, for permission to publish this article, and to Mr. G. Clayson for his assistance in the designing of the plant.

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The Radio Astronomy Observatories at Jodrell Bank and Cambridge

WHEN the 100in Mount Wilson telescope started its work just after the First World War, astronomers believed that the entire universe was continued within the confines of the Milky Way—a system containing millions of stars scattered throughout a roughly spherical enclosure across which light would take about twenty thousand years to travel. Moreover, it was firmly believed that the sun and solar system were at the centre of this great assemblage. The light-gathering power and penetration of the Mount Wilson telescope were so superior to any existing telescope, that it was soon realized that the Milky Way system was built on an altogether more gigantic scale than had been believed. Instead of a spherical enclosure the Milky Way system was revealed to be flattened with an extension of nearly one hundred thousand light years, but only a few thousand light years thick at the centre. This system is now known to contain something between one thousand million and ten thousand million stars, with the sun and the solar system much nearer the extremity than the centre.

This has been carried out with telescopes and other instruments receiving light waves emitted by the stars in the visual part of the electromagnetic spectrum. Auxiliary instruments, such as photo-electric cells and photographic plates, can extend these studies somewhat beyond the visual limits into the infra-red and ultra-violet regions, but appreciable extension is impossible because of the absorption caused by water vapour and fine dust in the earth's atmosphere. The distribution of the energy in a hot body such as a star is well known and it has seemed clear that knowledge of the universe was not appreciably restricted by this absorption in the atmosphere. This belief seemed so well founded that there was little astronomical interest in a second, more extensive gap or window in the atmosphere at very much longer wavelengths.

This other gap exists in the radio wave region. At its short-wave end, it is again limited by atmospheric absorption near a wavelength of a few centimetres and at the long-wave end by reflection in the Heaviside layer or ionosphere at a wavelength of about twenty metres.

The Discovery of Radio Waves from Space

At the end of 1931 an American engineer, Jansky, made the discovery that radio waves apparently emanating from regions beyond the solar system were reaching the earth through this window in the atmosphere, but it seems doubtful whether many astronomers knew of his work and the only important additions to his results before the Second World War were obtained by Grote Reber, an amateur investigator who built apparatus of advanced design in the garden of his home in Illinois. He found that the radio signals were strongest from directions near the centre of the Milky Way, and that the radio signals were roughly proportional in strength to the concentration of stars in the direction to which the radio telescope was pointing. On the other hand, Reber failed completely to detect any signals from the bright stars or from other prominent features visible in telescopes.

This paradox led him to the view that the radio signals were being generated in the very rarified hydrogen gas which fills interstellar space. This represented the extent of our knowledge of these radio waves from space in 1945. The Second World War placed in the hands of astronomers a new and enormously powerful tool for the exploration of space. The concentration of work on radio and radar for military purposes resulted in advanced techniques and when these were applied to the investigation of these radio waves generated in the cosmos spectacular results were obtained.

The Discovery of Radio Stars

The first measurements confirmed Reber's results and there seemed to be no direct connexion between the radio signals and the astronomical objects which comprise the universe familar to the human senses. Reber's idea, that the emissions were generated in the interstellar gas, remained for some time the only realistic suggestion, but in 1948 came the first of a sequence of discoveries which stimulated the interest of astronomers throughout the world. Bolton and Stanley in Sydney, followed immediately by Ryle and Smith in Cambridge, found that at least some of the radio waves were coming from localized sources in space, subsequently called radio stars. The two most intense of these sources were in the constellations of Cygnus and Cassiopeia. If these radio sources had coincided with any prominent visual objects the discovery would not, perhaps, have occasioned much surprise, but although both lay in densely populated stellar regions there were no particular visual objects to which the radio emissions could be attributed.

Subsequently many other, less intense, radio stars were discovered and there seemed to be no correlation with any class of the star known to astronomers; neither did any of the common stars appear to emit radio waves which could be detected on the earth. The belief arose that we were dealing with a new type of body in the heavens, dark or only faintly luminous but with the facility of emitting powerful radio waves; moreover, a type which appeared to be of frequent occurrence and distributed throughout the Galaxy in a manner similar to that of the common stars. For some time there was uncertainty as to whether the other galaxies might be similarly endowed with the facility of emitting intense radio waves, but any such doubts were laid to rest in 1950, when scientists at Jodrell Bank used a very large radio telescope and showed that the nearest galaxy in Andromeda behaved in a manner similar to our Galaxy, as far as emission of radio waves was concerned. Subsequently the emissions from many more remote galaxies or nebulæ were detected, and it is now widely accepted that the type of radio source responsible for the emission in our Galaxy, the Milky Way system, must be widely dispersed throughout the extragalactic star systems which comprise the universe.

The Identification of Radio Stars

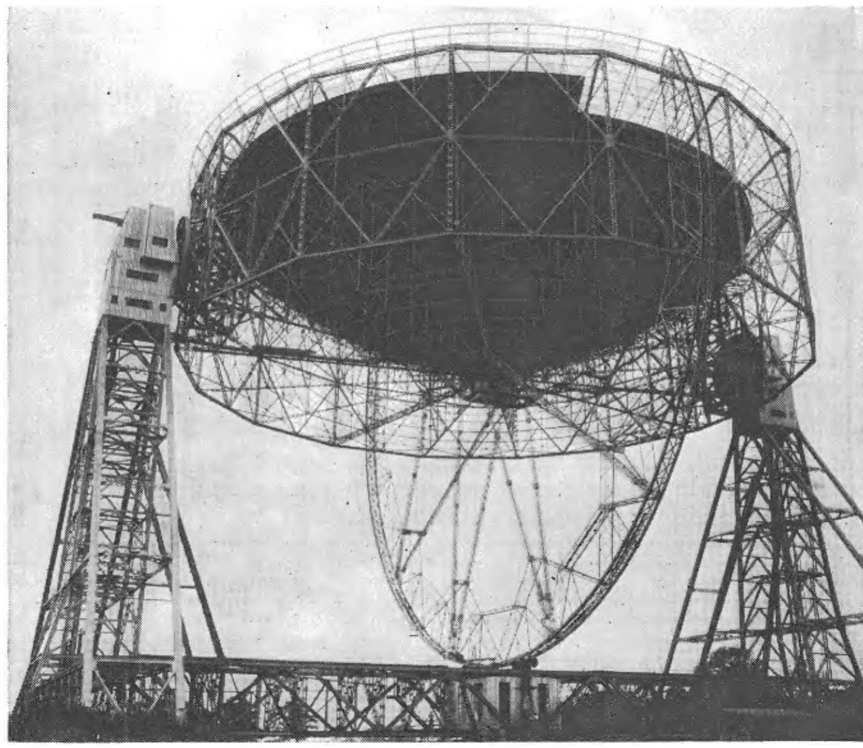
In the last few years there has been very close co-operation between the scientists using the radio telescopes and the astronomers with the big optical telescopes in America, in an effort to find a more precise relationship between the radio sources and objects which are visible in the telescopes. Although nearly two thousand of the radio sources have now been positioned, and in many cases the size and shape measured, the linkages which have been established with the common stars remain remarkably few, and the general paradox of the existence of the radio sources remains. It is, however, now certain there is one connexion which was suspected several years ago.

Occasionally a star blows up—it becomes a supernova. The atomic processes which generate the energy inside the star get out of hand and nature gives a replica on a really gigantic scale of an atomic bomb explosion. Instead, however, of an explosion of a few pounds of uranium or plutonium, all the millions of tons of material in the star blow up. Only three such explosions of stars in the Milky Way have ever been recorded. The most famous of these occurred a thousand years ago and the remains of the star can still be seen as an enormous cloud of tremendously hot gas travelling into space at a rate of seventy million miles per day. This object—the Crab Nebula—has now definitely been established as a source of strong radio emissions. In fact, it is the third strongest source in the heavens. The other two supernovæ—those discovered by Tycho Brahe in 1572 and by Kepler in 1604 are also known to be radio sources, although very much weaker. Although these connexions are of extreme interest, it seems unlikely that supernovæ can account for more than a few of the radio sources in the Galaxy.

A few years ago, scientists in Cambridge measured the position of the most intense source in Cassiopeia with such precision that it became possible to carry out extensive photography of the sky with the great Palomar telescope. In the position of this radio source, the Palomar telescope has photographed a peculiar object which is certainly not a star as commonly understood. It appears as a faintly luminous cloud of gas, spread out over a distance which is large compared with a star. Some of the gas is in extremely violent motion, and it seems likely that the generation of the radio signals must be connected in some way with this motion. Recently, two or three other similar objects have been located in the position of less intense radio stars, and there is also a good deal of speculation that this type of diffuse gaseous agglomeration may be responsible for many of the radio stars. There is no answer to the question as to the place of this type of object in the sequence of stellar evolution. The objects are so faint that they can only be seen by taking long exposures with the world's largest telescope, and yet they generate powerful radio signals. Some attempt has been made to link them with very old supernovæ, of an age such that they are not in the epoch of recorded history. On the other hand, they may equally well be stars at the opposite end of creation, that is those which are in the very early stages of formation.

The Collision of Nebulæ

Perhaps the most remarkable identification so far made is that of the second most intense radio source in the sky, which lies in the constellation of Cygnus. The early efforts to link up this radio source with a visible object, by inspecting the star maps, led to no result. There was nothing visible in the sky which seemed likely to be responsible for such a strong source of radio waves. About two years ago, however, as a result of the precision Cambridge measurements, referred to above, it again became possible for the American astronomers to train the Palomar telescope on the precise region of the sky which contained this radio source. Their plates yielded the usual large number of faint stars and nebulæ, but in the position of the radio source there was an object with an unusual appearance. It was interpreted as showing two great galaxies in a state of collision. The distance of this celestial collision in Cygnus is enormous—a hundred million light years or just about the limit of the present observable universe.



The partly constructed radio telescope at Jodrell Bank

The establishment of the connexion between this faint object at the limit of the observable universe, with the powerful source of radio emission in Cygnus, carries with it one of the most surprising implications in the whole of radio astronomy. The study of the radio records, coupled with the data about the distribution of these extragalactic nebulæ in space, leads one to conclude that this type of celestial collision is by no means unique, and that it might even be a fairly common event. In the space around the Milky Way system, the distance between nebulæ is of the order of a million light years. This is a fairly average distance between nebulæ in space and the chances of collision are negligible. On the other hand, in some regions of space the galaxies are far more closely packed in clusters, and in these one finds, perhaps a thousand galaxies separated by only thirty thousand light years. Although this distance is still very great, the galaxies themselves are moving with a speed of about fifteen hundred miles per second and a simple calculation indicates that the chances of a collision are considerable. Now in the case of the Cygnus collision a powerful radio signal can be received although the two colliding nebulæ are so distant that they are just at the limit of the universe observable by the big telescopes. If the galaxies were even further away, so that they could just be identified by the telescopes, it would still be possible to detect their radio emission, even with the present radio telescopes. In fact, it is fairly easy to calculate that even with the sensitivities of present radio equipment, the colliding galaxies in Cygnus could be two or three times further away and still be measurable with the radio telescopes. Hence, if this type of celestial collision is as common as is now believed, even our present radio telescopes are capable of studying a volume of space which is, perhaps, five or ten times greater than the volume which is accessible to the greatest optical telescope in the world. This kind of possibility will almost certainly be greatly enhanced when the new radio telescopes now under construction at Jodrell Bank and Cambridge come into operation.

The Spectral Line Emission from the Neutral Hydrogen Gas

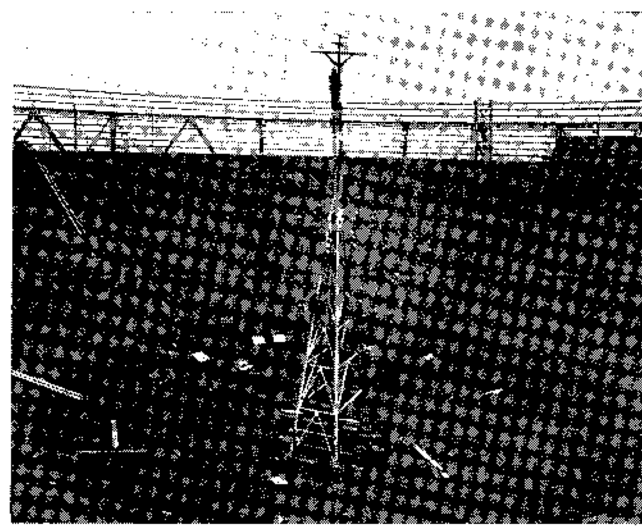
The radio emissions from space are emitted over a wide range of wavelengths. Although the precise nature of the spectrum has not yet been established, it is well known that these radio sources can be detected over a range of wavelengths from a few centimetres to 15 or 20 metres. During the last year or so, however, a great deal of attention has been given to another type of radio emission from space, which is generated in the neutral hydrogen gas in the Milky Way on a wavelength of 21 centimetres. This is a spectral line, which is emitted when the spin of the electron in the ground state of a neutral hydrogen atom reverses. The possibility that this type of radio emission might occur was first suggested by van de Hulst in 1944 but it was not until 1951 that scientists in America, Australia and Holland succeeded in detecting and measureing this emission. The detection of this spectral line is in itself a remarkable achievement. The hydrogen clouds are tens of thousands of light years distant and contain only about one hydrogen atom per cubic centimetre. Moreover, this change of spin in the atom is only likely to happen once in about eleven million years. It seems, however, that the atoms in their random motions collide every fifty years, and in a collision there is a one in eight chance that this transition will occur.

The success in detecting this line in 1951 immediately opened up important new possibilities in radio astronomy. The hydrogen clouds are in motion relative to the solar system and this 21 centimetre spectral line will, in consequence, show a doppler shift in its frequency. Using this technique astronomers in Leiden have been able to study the detailed motions of the hydrogen clouds in the region of the Milky Way, which are obscured from the view of the optical telescopes by the great dust clouds.

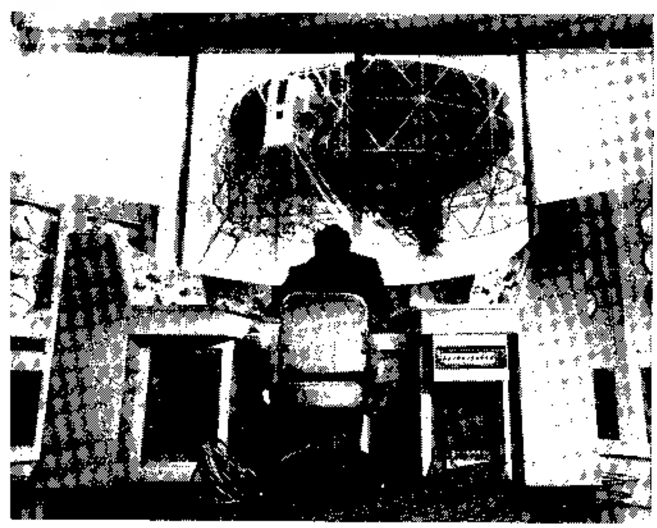
The Need for Large Radio Telescopes

These studies of the radio emissions from space are carried out on a wavelength a million times longer than the wavelength of the light waves which are focused by the conventional optical telescopes. The radio waves are unaffected by cloud, fog or daylight and in this respect the radio astronomers has a marked advantage over the traditional methods of astronomical investigation. On the other hand, because of the long wavelength, it is extremely difficult to achieve any appreciable resolution. The beam width, or the angle of the cone in which the radiation is received, depends on the ratio of the wavelength to the diameter of the telescope. Thus, to achieve the same resolution as a very small optical telescope, the aerials of a radio telescope would have to extend for thousands of miles. The need for the maximum possible resolution in the radio work has been a dominant feature of the technical developments. A great deal has been achieved by special devices in which two similar aerial systems, spaced by several hundred yards, are connected to a common recording equipment. This type of radio telescope, known as an interferometer, has been intensively developed in Cambridge and Sydney, and it was with systems of this type that the original discovery of radio stars was made.

In an alternative approach, the physical size of the aerial system is increased. There are now several steerable radio telescope of small size in existence and the largest is believed to be the transit radio telescope at Jodrell



The aerials of the radio telescope under construction



A view inside the control room, looking out on the telescope

Bank, which has an aperture of 218ft, but this is fixed to the earth and only a small part of the heavens can be explored. Experience with this instrument soon demonstrated that a completely steerable radio telescope of this order of size was a prerequisite for the further exploration of space by the radio method. The engineering difficulties and expense of such an undertaking are formidable. Nevertheless, the results to be anticipated were such as to enlist the sympathetic interest of many prominent scientists, when the idea was first put forward in 1949.

In conventional astronomical investigations, the telescope has been the crucial instrument for the exploration of space. Successive increases in size have led to more light-gathering power and greater resolution and although the improvements in auxiliary instruments, such as photographic plates and spectroscopes, have been very important, nevertheless the great advances in observational astronomy have come primarily from larger and larger telescopes. The situation in radio observations is very similar. Whereas large optical telescopes are required to improve the light-gathering power and the resolution, large radio telescopes are required in order to be able to pick up faint signals at greater distances from the earth, as well as for greater resolution. The Department of Scientific and Industrial Research and the Nuffield Foundation gave financial backing to this proposal for a large steerable radio telescope. This telescope nearing completion at the Jodrell Bank Experimental Station of the University of Manchester, will have a paraboloidal reflector of steel sheet with an aperture of 250ft. This moving bowl will weigh 750 tons and by supporting it on steel towers rising 180ft above the ground, it will be possible to direct it to any part of the sky. The foundations contain thousands of tons of reinforced concrete to hold the 17-foot gauge double railway track on which the telescope rotates. The steel towers run on 12 bogies driven by electric motors, and the racks for the elevation movement have been taken from the dismantled battleship *Royal Sovereign*. There are grounds for hope that preliminary tests with this telescope in a fixed position may be made in the autumn of 1957 and that the telescope will be under full powered control in 1958.

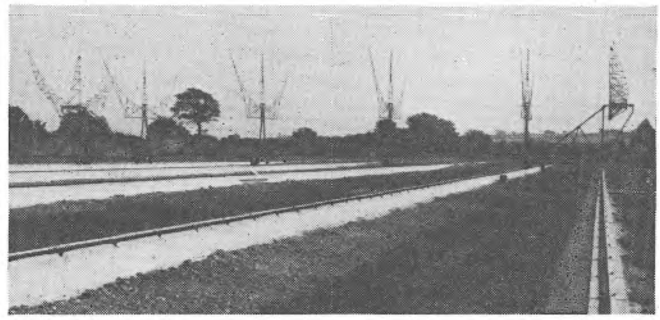
The beamwidth of this telescope will be about one degree when used on a wavelength of one metre and its power gain will be over sixteen thousand. On the wavelength of the spectral line at 21 centimetres, its beamwidth will be a few minutes of arc only. The great discrimination and the power gain will be used to elucidate the nature of the radio sources and to plot detailed maps of the sky, particularly in those regions which are obscured by the interstellar dust. In the field of the 21 centimetre emission great importance is attached to the extension of these studies to the extragalactic nebulæ. At the moment, it seems likely that the fascination and importance of these problems associated with the radio emissions from space will occupy nearly the whole time of the telescope for the first few years of its working life. On the other hand it has always been intended that the radio telescope should be used in all aspects of radio astronomy, including those in which radio waves are first transmitted from the earth.

When the telescope is being used to receive radiations from outer space the amount of energy concentrated on the aerial at the focal points is extremely small. In order to amplify and analyse these radiations it is desirable that the radio receiving equipment should be located as close as possible to the aerial. For some purposes the first amplifier will be immediately adjacent to the aerial at the summit of the 66ft tower in the centre of the bowl, but the main amplifiers are suspended in a steel laboratory hung immediately below the centre of the bowl itself. This laboratory is suspended on trunnions in order to keep its floor level and is equipped with damping devices to prevent uncomfortable oscillations due to the high winds to which it will be subjected. The suspended laboratory can be reached by gangways from the top of each bearing tower, where further high level laboratories travel round with the azimuth motion of the telescope. Each tower contains an electrical passenger lift and also a hoist for handling mechanical equipment to the motor rooms.

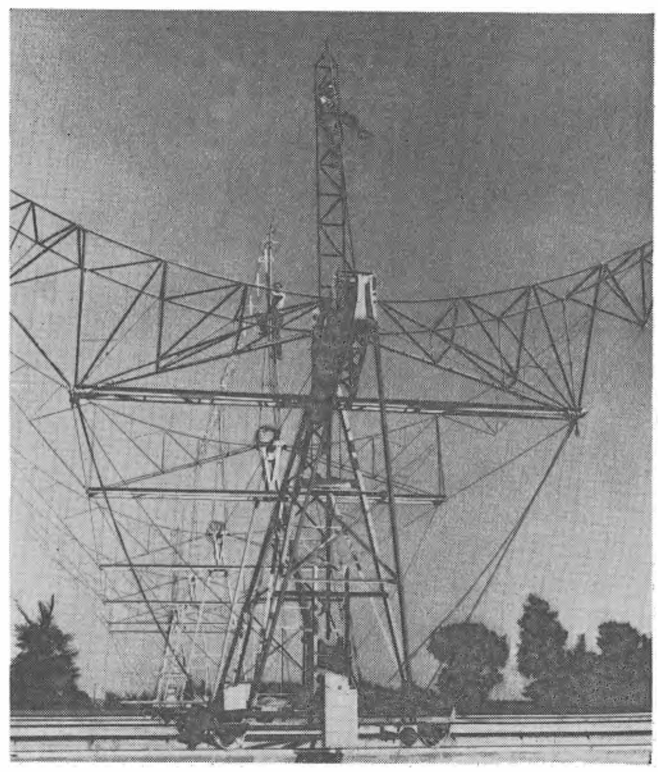
A control desk and computer mechanism are housed in a building some 200 yards from the centre of the telescope. The control system and computer permit the telescope to follow automatically any point in space irrespective of the rotation of the earth and the movement of the earth round the sun. Facilities are also provided for scanning predetermined areas in space and for tracking moving objects.

Radio Astronomy at Cambridge

Radio astronomy was started soon after the war by Martin Ryle in the Cavendish Laboratory at Cambridge. The work was primarily concerned with the detection of radio stars and for this purpose an 'interferometer' type of aerial system was developed. Its method of working can be understood as follows. The 'resolving power' of an aerial depends on the interaction or 'interference' of the



The moving aerial and railway tracks of the interferometer

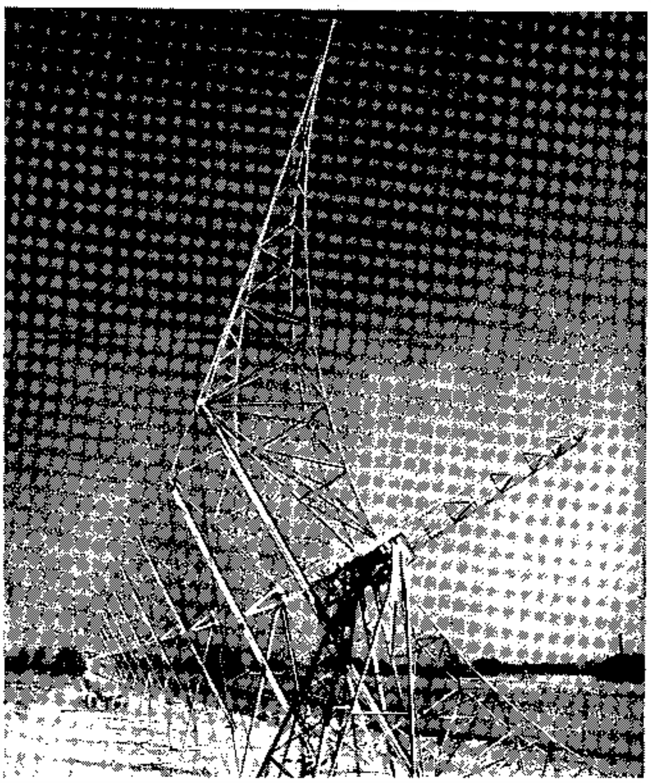


A close up view of one of the aerial towers of the interferometer

radio waves received at its edges, and the further apart these edges are, the greater is the resolving power. Suppose therefore that an aerial is made as large as possible, so that because of its large area it collects a large power and because of the large distance between its edges it has great resolving power. Now suppose it to be split into two halves which are then moved apart some considerable distance. Then the total power received is the same as before; but if the waves received in the two halves are properly combined they will, by their 'interference' effect, provide a resolving power which is increased proportionally to the distance between them. In this way it is possible to increase the resolving power far beyond what is practicable with a single large aerial, and at the same time to keep the collecting power the same.

An interferometer aerial of great resolving power was constructed at the Cavendish Laboratory, Cambridge in 1948 to make a survey of the distribution of radio stars over the sky.

In 1952 a new and much larger radio telescope of the interferometer type was built, with the aid of a grant from the Department of Scientific and Industrial Research. The total collecting area of this is equal to the collecting area of the large single aerial constructed at Manchester, and its resolving power is that appropriate to the distance, 1 900ft, between its two halves. About 2 000 radio stars have been detected with this instrument but only about ten of these can be identified with visual objects.



The fixed aerial of the interferometer

Plans were made in Cambridge to extend the observations by constructing an even larger 'interferometer' in which both the collecting power and the resolving power would be increased. This was too large to be put on the same site as the existing interferometer and it was necessary to plan in terms of a new radio astronomy observatory covering an area large enough to contain both this, and other large aerials.

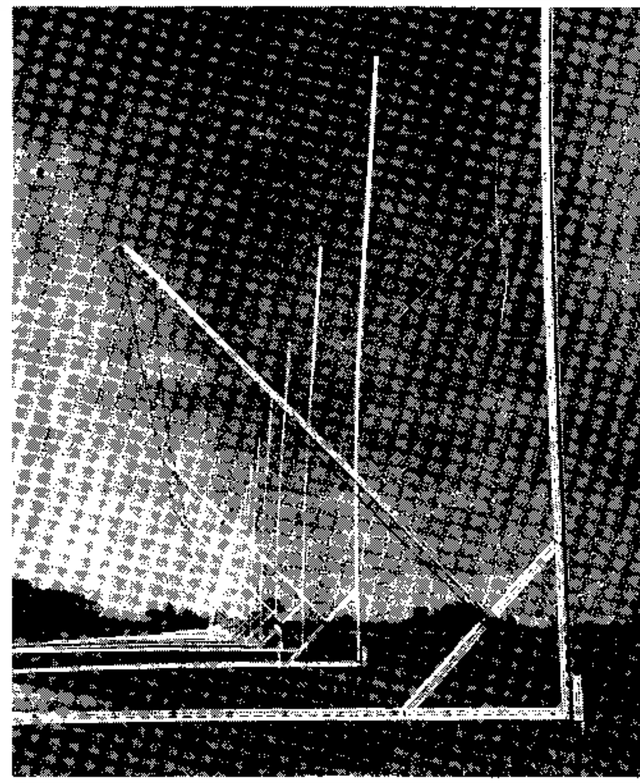
In 1955 The University of Cambridge received from the Mullard Company an offer to provide over a period of ten years the sum of £100 000 for the purpose of continuing and extending the work in radio astronomy which is being done by the Cavendish Laboratory.

With this benefactor together with a grant from the D.S.I.R. a new Radio Astronomy Observatory has been established at Lords' Bridge about five miles from Cambridge.

A small observatory building contains the radio equipment and recording apparatus and the new interferometer aerial system is now nearing completion.

The interferometer, which is designed for a wavelength of 1·7m, comprises a fixed east-west aerial 1 450ft long and 65ft wide; the moving aerial is 190ft long in the east-west direction and 65ft wide, and moves along north-south railway lines which are 1 000ft in length. Both aerials are cylindrical parabolas made as in the earlier Cambridge interferometer by stretching wires across parabolic tubular steel frames which can rotate about an east-west axis.

The resulting envelope pattern will have a width to half intensity of approximately 25 by 35min of arc. The equivalent 'collecting area' of the instrument will be approximately 190 000ft^2, which will provide considerably greater sensitivity than any previous radio-star instrument.



The fixed aerial of the 'pencil beam' system

The 'pencil beam' system for galactic study is designed for a wavelength of 7·9m, and uses a fixed east-west aerial 3 200ft long and a moving aerial which has a range of 1 700ft. Each aerial is in the form of a corner reflector having an aperture of 40ft. The reception pattern of the instrument will be approximately 1° by 1° to half intensity, and the equivalent collecting area is approximately 200 000ft^2.

The operation of both instruments will involve a large amount of mathematical computation to synthesize the recorded data. The co-operation of the University Mathematical Laboratory has been sought, and methods have already been developed for the synthesis on EDSAC of the records obtained with the earlier instrument; similar methods will be employed for the new instruments.

In addition to the two large radio telescopes which are now being built, the area of level ground which is available will make possible many investigations employing simpler systems of high resolving power; it is hoped in this way to extend a number of lines of work which have been investigated at the present site.

A Low Voltage Stabilizer

Employing Junction Transistors and a Silicon Junction Reference Diode

By D. Aspinall*, B.Sc.

A series voltage stabilizer using junction transistors is described. A Zener diode is used as a reference voltage source. Approximate equations for the output imedance and stabilizing factor of the circuit are obtained and found to check with experiment. The change of output impedance with frequency is also investigated experimentally.

THE voltage supplies for large transistor equipments have been provided mainly by accumulators. This source of supply has good regulation properties and is adequate except for its need of maintenance. The regulation properties of the simple rectifier and subsequent filters make it suitable for only small currents, and so there is a need for some form of easily regulated small voltage and high current supply.

Power transistors which have recently become available in this country, make it possible to design stabilizers with a current handling range from 0 to 1A at voltages in the region of 10V. Transistor stabilizers have also been considered in the United States, where power transistors have been available for some time[1,2]. Junction transistors currently available from British manufacturers have been incorporated into the series stabilizer which is now described.

The basis of any voltage stabilizer is a source of constant voltage, which is used as a reference. Neon tubes are frequently used in thermionic valve stabilizers, but the relatively high working voltage of these devices precludes their choice in a transistor stabilizer. Reference voltage cells can be manufactured, based upon the constancy of the electrochemical potential. A particular cell of Belgian manufacture has the following characteristics: when a current is passed through the cell a potential difference of 1·5V is developed across the terminals which is largely independent of changes in current. (Current change 100 to 300mA—voltage change 1·45 to 1·5V.)

Use of these cells is a possibility, but certain silicon diodes offer more favourable characteristics. Such Zener diodes, as they are termed, when biased in the reverse direction, can be used as constant voltage sources. The use of these diodes will now be discussed in greater detail.

Reference Voltage

The reverse current of a silicon diode is small until a reverse voltage termed the saturation or Zener voltage, is reached. At this voltage the current carriers are given sufficient energy to form further current carriers; a cumulative action results and it is possible for a high current to flow. The voltage at which the saturation occurs is controlled during manufacture.

A diode is available in this country which has a saturation voltage of -7V, and its characteristic is shown in Fig. 1. It can be seen that the resistance of the diode during saturation is very low and of the order of 1·5Ω.

The maximum current which can flow through the diode is limited in practice by the heat generated at the junction, the temperature of which must not rise above some critical temperature which in the case of the diode whose characteristic appears in Fig. 1 is 150°C. A graph of maximum input power against ambient temperature is shown in Fig. 2. As the voltage across the device is constant, it is convenient to plot maximum current against ambient temperature. The highest current which can be drawn at room temperature is approximately 100mA. Thus the direct use of such Zener diodes as voltage stabilizers is strictly limited

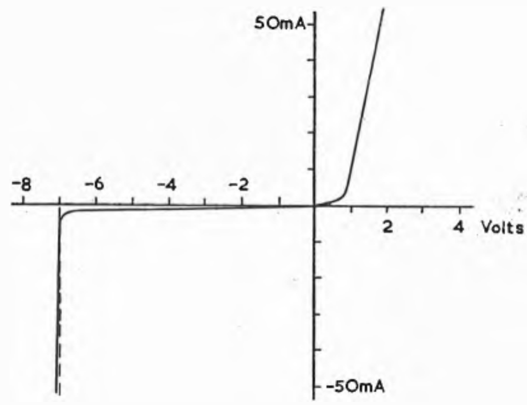


Fig. 1. Characteristic of B.T.H. type STDX3/7. Silicon junction diode

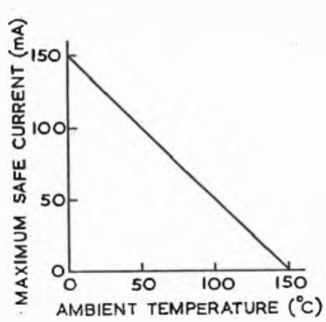


Fig. 2. Ambient temperature V_e. Max. power dissipation

because of the restricted current handling properties. It is, however, feasible to use these diodes in voltage stabilizers containing current amplification.

The Series Stabilizer

Fig. 3 shows the basic design of the stabilizer in which a power transistor (X_2) is placed in series with the supply voltage and load resistance. This transistor can be considered as a series variable resistor which is controlled by the output voltage V_o amplified by the parallel or control transistor (X_1). A reference voltage is provided by the Zener silicon diode.

The emitter voltage of X_1 will be constant with respect to the positive rail due to the reference diode. This diode maintains a constant voltage across its terminals, provided the current flowing through it is sufficient to maintain saturation. The voltage V_1 between the base of X_1 and the

* *The University, Manchester.*

positive rail will be the sum of the reference voltage across the diode plus the emitter-base voltage of X_1 The output voltage will be $V_1(R_1 + R_2)/R_1$ and the emitter-base voltage of X_2 will define the collector voltage of X_1. Care must be taken to ensure the collector dissipation of X_1 is not exceeded when deciding upon an actual value of output voltage. The current flowing through R will depend upon E and the base voltage of X_2. During the operation of the stabilizer this base voltage and E may vary (base voltage change less than change in E). E must therefore be made large to swamp these variations.

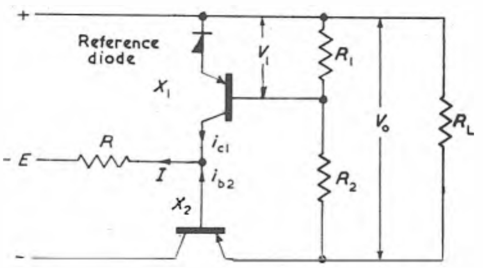


Fig. 3. Series stabilizer

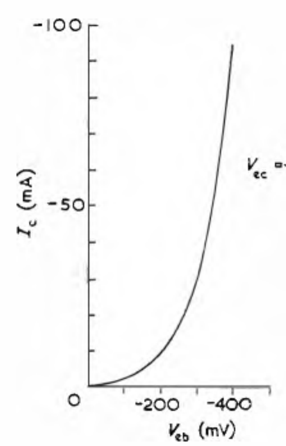


Fig. 4. $V_{eb}-I_c$ characteristic Mullard type OC72

The current I is made up of collector current in X_1, (i_{c1}) and base current in $X_2(i_{b2})$.

$$I = i_{c1} + i_{b2}$$

When the output of the stabilizer is open-circuited, ($R_L = \infty$), i_{b2} will be quite low since the only emitter current flowing in X_2 is that due to the presence of the potentiometer chain. Assume that when a load current is taken, by making R_L finite, the output voltage falls by δV_o. This implies that the voltage V_1 also falls by

$$(\delta V_o R_1)/(R_1 + R_2)$$

Since the potential of X_1 emitter is constant, the base-emitter voltage of X_1 decreases and this leads to a drop in the collector current flowing in X_1. The current I is sensibly constant so that the reduction in i_{c1} must lead to an increase in i_{b2}. The current gain of X_2 causes the emitter current in X_2 to increase, thus raising the output voltage, ideally by δV_o.

Perfect stabilization cannot be produced since some change in output voltage must occur before a correction is applied. The action depends to a large extent upon the mutual conductance (g_m) of X_1.

$$g_m = I_c/V_{eb}$$

g_m must be high so that a small change in emitter-base voltage leads to a large change in collector current. Fig. 4 gives the characteristic of the best transistor currently available. It has a high g_m (of the order of 0·6 mho), for high collector currents, but below about 20mA the g_m falls rapidly. If this transistor is to be used as X_1 in Fig. 3, it will be necessary to ensure the collector current never falls below 20mA.

The stabilization against changes in input voltage will now be considered. The allowed change in input voltage is quite large and only depends upon the characteristics of X_2. The junction transistor approximates to a constant current device if the collector-emitter voltage is above a certain minimum, the bottoming value. The voltage across X_2 can therefore vary over very wide limits provided the dissipation is not exceeded on the upper excursion and the transistor does not bottom on the lower.

The output will, however, be sensitive to changes in the defining voltage E. If this voltage changes, the current in R is altered and the collector current of X_1 and the base current of X_2 will be modified, leading to a change in output voltage. The effect of variations of E on the output voltage will now be taken into consideration.

Stabilization Against Changes in Defining Voltage E

Referring to Fig. 3. Assume the voltage E varies by δE leading to a change in output voltage δV_o.

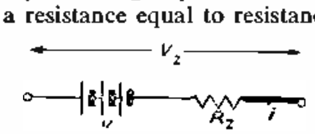
The change in emitter current in X_2 will be

$$i_{e2} = \delta V_o/R_L \text{ (assuming } R_L \ll R_1 + R_2)$$

$$i_{b2} = (1 - \alpha_2)\,\delta V_o/R_L \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

$$\alpha_2 = \text{current gain of } X_2.$$

This change in output voltage will produce a change in the emitter-base voltage of X_1 which will in turn bring about a change in i_{c1}. The change of voltage at the base δV_1 will be made up of the change in v_{eb} and the change in emitter voltage due to the finite resistance of the reference diode. A simple equivalent circuit of the Zener diode is a battery of voltage equal to the saturation voltage in series with a resistance equal to resistance at saturation.



$$V_z = V_s + R_z i$$

This equivalent circuit is only valid so long as the current flowing is sufficient to maintain saturation.

$$\delta V_1 = \delta V_z + \delta V_{eb1} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

$\delta V_z = R_z\,\delta i_{c1}/\alpha_1$ $\quad$ R_z = Resistance of reference diode

$\delta V_{eb1} = \delta i_{c1}/g_{m1}$ $\quad$ α_1 = Current gain of X_1

g_{m1} = Mutual conductance of X_1

Equation (2) now reads

$$\delta V_1 = [(R_z/\alpha_1) + (1/g_{M1})]\delta i_{c1}$$

but $\delta V_1 = (1/\beta)\delta V_o$

$$\beta = (R_1 + R_2)/R_1$$

$$\therefore\ \delta V_o = \beta\,[(R_z/\alpha_1) + (1/g_{M1})]\,\delta i_{c1} \ldots\ldots \quad (3)$$

The current I will not be constant due to the changes in E, V_o, and V_{eb2}.

Let change in $I = \delta I$

$$\delta I = (1/R)(\delta E - \delta V_o - \delta V_{eb2})$$

Now $\delta V_{eb2} = (\delta i_{e2}/g_{m2})$

g_{m2} = mutual conductance of X_2

also $\delta i_{e2} = (\delta V_o/R_L)$

$$\therefore\ \delta I = (1/R)(\delta E - \delta V_o - (\delta V_o/\alpha_2 g_{m2} R_L)) \quad \ldots \quad (4)$$

Now $I = i_{c1} + i_{b2}$

$$\therefore\ \delta I = \delta i_{c1} + \delta i_{b2} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (5)$$

Substituting in equation (5) from equations (1), (3) and (4)

$$\frac{1}{R}\left(\delta E - \delta V_o - \frac{\delta V_o}{\alpha_2 g_{m2} R_L}\right) = \frac{\delta V_o}{\beta[(R_z/\alpha_1) + (1/g_{m1})]} + \frac{(1-\alpha_2)\,\delta V_o}{R_L}$$

$$\therefore \frac{\delta E}{\delta V_o} = 1 + \frac{\alpha_1 g_{M1} R}{\beta(g_{m1}R_z + \alpha_1)} - \frac{R}{R_L}(1-\alpha_2) + \frac{1}{R_L \alpha_2 g_{m2}}$$

Call $(\delta V_o/\delta E)$ the stabilizing factor.

Inserting typical values $(\delta V_o/\delta E)$ equals $1/40$.

To a first approximation

$$\frac{\delta V_o}{\delta E} = \frac{\beta(g_{m1}R_z + \alpha_1)}{\alpha_1 g_{m1} R} \qquad \text{.............} \qquad (6)$$

Output Impedance of the Stabilizer

The other important feature which must be considered is the change in output voltage with a change in load current; the output impedance of the stabilizer.

$$I = i_{b2} + i_{c1}$$

$$I = i_{e2}(1-\alpha_2) + i_{c1} \qquad \text{....................} \qquad (7)$$

$$i_{c1} = g_{m1}V_{eb} = g_{m1}((V_{out}/\beta) - R_z i_{e1} - V_s)$$

$$= g_{m1}((V_{out}/\beta) - R_z(i_{c1}/\alpha_1) - V_s)$$

$$\therefore i_{c1} = \frac{g_{m1}V_{out}}{(1 + (g_{m1}R_z/\alpha_1))\beta} - \frac{V_s g_{m1}}{1 + (g_{m1}R_z/\alpha_1)} \qquad \text{......} \qquad (8)$$

Substituting in equation (7) from equation (8)

$$I = i_{e2}(1-\alpha_2) + \frac{g_{m1}V_{out}}{(1 + (g_{m1}R_z/\alpha_1))\beta} - \frac{V_s g_{m1}}{1 + (g_{m1}R_z/\alpha_1)}$$

$$\therefore \quad \frac{\delta V_{out}}{\delta i_{e2}} = -(1-\alpha_2)(1 + (R_z g_{m1}/\alpha_1))(\beta/g_{m1})$$

Output impedance =

$$(1-\alpha_2)(1 + (R_z g_{m1}/\alpha_1))(\beta/g_{m1})$$

and inserting values = 0·2Ω.

Design Consideration of a Practical Circuit

RANGE OF OPERATION

A practical circuit is drawn in Fig. 5. The operating range of this stabilizer will be from zero load current to some maximum current which will be governed by the limitations of the components used in the stabilizer. The maximum current is finally governed by the maximum emitter current which X_2 can safely pass. A limit may, however, be reached before this, due to the limit which is imposed upon the base current in X_2 by the current in the defining resistor R. The maximum current which can flow in this resistor is governed by the maximum collector current which is to be allowed in X_1. This is in turn governed by the maximum dissipation of X_1 and the reference diode. Therefore the maximum base current flowing in X_2 cannot be greater than the maximum allowed collector current in X_1 less that current which must still be allowed to flow in X_1 to maintain the reference voltage and the g_m of X_1.

DESIGN OF POTENTIOMETER

Though the change in the emitter-base voltage of X_1 affects the control of current in X_1 and hence the stabilization of the output voltage; the finite input impedance of X_1 has to be considered when the potentiometer chain is being designed.

When the stabilizer is off load i_{e1} will be a maximum and therefore i_{b1} will also be a maximum. As the load is increased these two currents will fall. This results in a change of current flowing out of the base of X_1 and into the potentiometer chain which will change the value of V_1 and detract from the stabilization. The current flowing in the potentiometer chain must therefore be quite high to swamp these changes in base current

i_{e1} max. = 70mA. $\qquad$ i_{b1} max. = 1·4mA.

i_{e1} min. = 20mA. $\qquad$ i_{b1} min. = 0·4mA.

∴ current in chain ≫ 1mA.

and in practice 100mA is used.

CHOICE OF INPUT VOLTAGE

The choice of input voltage will depend upon the particular output voltage. It must be large so that R can be large, but it cannot be too large otherwise the dissipation

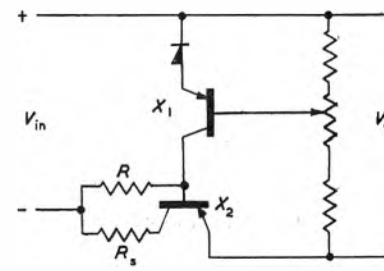


Fig. 5. Practical circuit

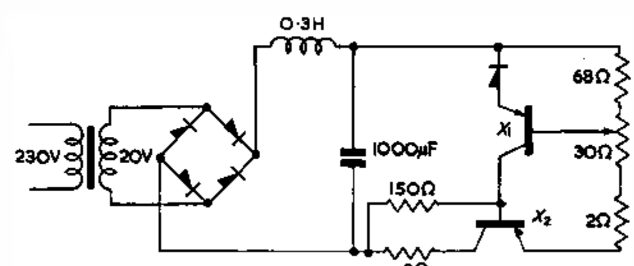


Fig. 6. Experimental circuit

in X_2 will be exceeded. The safety resistor R_s must be considered in this respect, see Fig. 5.

Power dissipation of $X_2 = P$

$$P = (V_{in} - V_{out} - R_s i_{c2})i_{c2}$$

$$P_{max} = \frac{(V_{in} - V_{out})^2}{4R_S} \qquad \text{..................} \qquad (9)$$

Equation (9) contains the relation between V_{in} and R_s but a further equation must be found to be able to deduce the actual values of R_s and V_{in}.

The emitter-collector voltage of $X_2 = V_{ce2}$

also $V_{ce2} = V_{in} - V_{out} - R_s i_{c2}$

As the collector current increases this voltage decreases. Ultimately therefore X_2 will bottom. The current at which bottoming occurs must be chosen to lie beyond the limitations on current in X_2 imposed by the current in R. Calling this bottoming current i_{c2}'

$$V_{ce2(min)} = V_{in} - V_{out} - R_s i_{c2}' \qquad \text{.......} \qquad (10)$$

let $V_{in} - V_{out} = \Delta$

$$\frac{\Delta - V_{ce2(min)}}{i_{c2}'} = R_s$$

but

$$R_s = \frac{\Delta^2}{4P_{max}} \qquad \text{...............} \qquad (11)$$

$$\therefore \; i_{c2}'\Delta^2 - 4P_{max}\Delta + 4P_{max}V_{ce2(min)} = 0$$

the optimum value of Δ =

$$\frac{2P_{max} + 2\sqrt{[P_{max}^2 - P_{max}\,i_{c2}'\,V_{ce2(min)}]}}{i_{c2}'}$$

Having found Δ it is now possible to find R_e from equation (11).

Choice of Defining Resistor R

The input voltage having been decided by the limitations of the transistors, R must now be chosen to give sufficient current when the stabilizer is off load to provide the i_{b2} and i_{c1} to the limit of the dissipation of X_1 or the reference diode.

Experimental Circuit

A circuit was built to this design as shown in Fig. 6.

The input was from a bridge rectifier using G.E.C. type EW54 medium power pn junction diodes. The output from

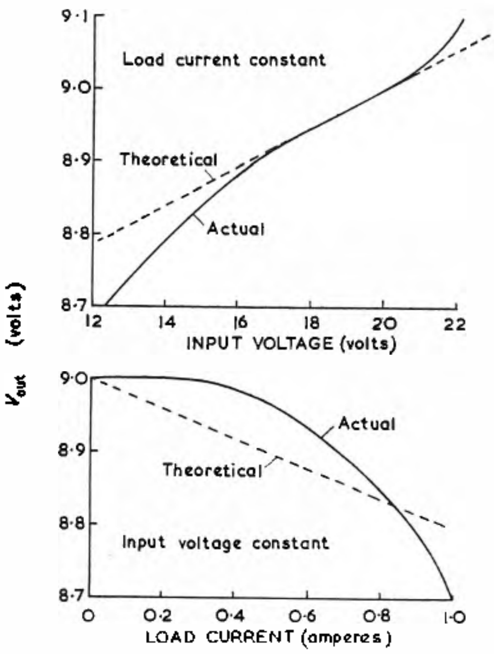


Fig. 7. Experimental results

the rectifier was smoothed by a 0·3H choke and 1 000μF capacitor, choke input filter. The control transistor X_1 was a Mullard type OC72 pnp junction transistor. This was chosen for its power handling properties and its eminently suitable $i_c - v_{eb}$ characteristic. The series transistor X_2 was a Pye type power V30/30P pnp junction transistor, chosen because it was the only power transistor available which could pass the relatively high currents required. The reference diode was an experimental silicon junction diode manufactured by the B.T.H. Research Laboratories type STDX3/7, also being the only one available.

The important properties are listed below:

OC72

$$\alpha = 0{\cdot}98$$

$$g_m = 0{\cdot}65 \text{ at } 70\text{mA}$$

$$= 0{\cdot}35 \text{ at } 30\text{mA}.$$

Maximum collector dissipation at room temperature $= 150$mW.

V30/30P

$\alpha = 0{\cdot}93$

Maximum collector dissipation at room temperature $= 4$W.

Maximum collector current $= 3$A.

STDX3/7

Saturation voltage $= -7$V

$$R_z = 1{\cdot}5\Omega$$

Maximum current at room temperature $= 100$mA.

It was decided to work the reference diode below its maximum current. The maximum current chosen was 75mA. The maximum collector current in X_1 is therefore of the order of 75mA.

The choice of output voltage depends largely upon the saturation voltage of the Zener diode. The minimum output voltage is this saturation voltage plus the emitter-base voltage of X_1. In this case $\beta = 1$ and the output impedance and stabilization factor are both at their optimum values.

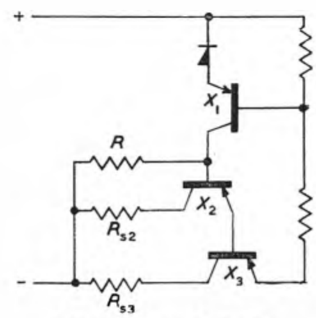


Fig. 8. Improved circuit

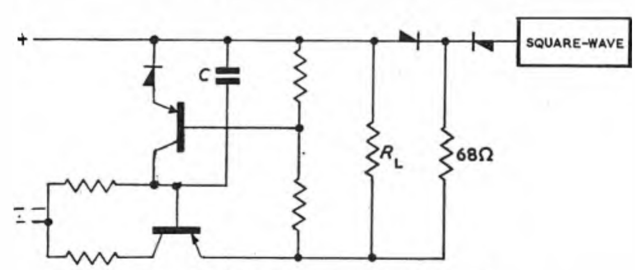


Fig. 9. Circuit to test transient response

As the output voltage is increased β increases and the quality of the stabilizer suffers. As the output voltage is increased the dissipation of X_1 must be watched carefully. If necessary a safety resistor may be placed between the collector of X_1 and the base of X_2 connecting R between the input and the base of X_2. In the particular example here considered the output voltage was fixed at 9V. This meant that the emitter-collector voltage of X_1 was in the region of 2V and the dissipation maintained within the manufacturer's limit.

The maximum base current which can flow out of X_2 is in the region of 75mA. Thus the maximum emitter current which can flow in X_2 is 1A at full load. Noting this fact and the maximum dissipation of X_2 the input voltage was fixed at 20V, with a safety resistor of 9Ω. The resistor R was chosen to be 150Ω.

This circuit maintained its stabilizing action from zero to 1A load current. Over this range the circuit was found to have an output impedance of the order of 0·3Ω and a stabilizing factor of 0·025. The results are shown in Fig. 7. They agree quite well with the theoretical predictions when the changes of the parameters with current and biasing values are taken into consideration.

An Improved Design

An improvement can be brought about by incorporating a further stage of gain in the feedback loop. An emitter-follower circuit as shown in Fig. 8 is found to be suitable.

In this circuit advantage has been taken of the fact that X_2 will amplify the collector current changes in X_1. This amplification enables the resistor R to be increased. This will improve the stabilization of output voltage against changes in input voltage but tends to increase the output impedance since the g_m of X_1 is reduced as its collector current is reduced. This is more than compensated by the current gain in X_2, which reduces the change of current in the collector circuit of X_1 and hence the change in current passing through the Zener diode. The reference voltage is therefore more constant. If the output impedance is deduced the following equation is obtained:—

$$\text{Output impedance} = (1-\alpha_2)(1-\alpha_3)(1+(R_2 g_{m1}/\alpha_1))(\beta/g_{m1})$$

When tested this circuit was found to have an output

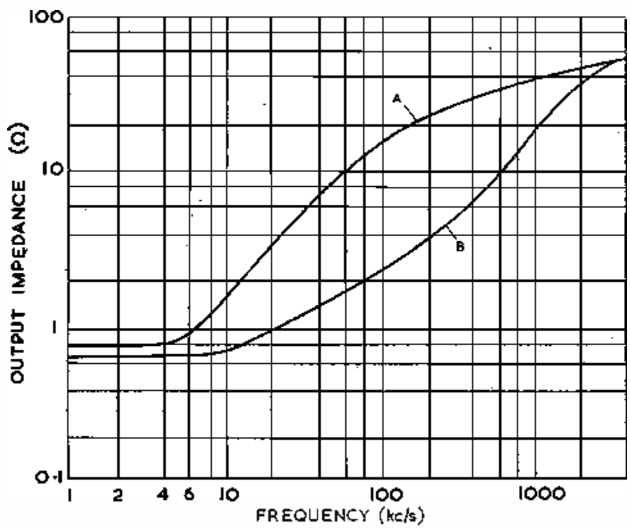


Fig. 10. Frequency response of stabilizer

impedance of 0·09Ω and a stabilizing factor of 0·0125 over a range of currents from zero to 1A. The safety resistors in the collector circuits of X_2 and X_3 must be carefully chosen to ensure the dissipations are not exceeded and that X_2 and X_3 bottom together.

Transient Response

The transient response of the stabilizer was tested by switching a current out of the stabilizer by means of the circuit shown in Fig. 9. G.E.C. type EW54 junction diodes were used because of their low forward impedance. It was found that a square wave appeared across the output due to the finite output impedance. There was also a considerable amount of overshoot due to the slow response of the transistor circuits. These overshoots were of 2·5V in magnitude, and 10 to 15μsec in duration. A capacitor was placed between the base of the series transistor and the positive rail. This capacitor by-passes the control transistor and provides current to the base of the series transistor during the transient. However, the limited frequency response of the series transistor still results in an overshoot but of reduced amplitude, being of the order of 0·5V.

A test of the frequency response was also carried out. The variation of output impedance with frequency is plotted in Fig. 10. Curve A is for the circuit without the feedback capacitor; curve B shows that when the feedback capacitor is included a slight improvement is brought about which is lost again at the highest frequencies. The low impedance of the stabilizer can be maintained at the higher frequencies by placing a capacitor across the output terminals, the reactance of which will decrease with frequency; e.g. a capacitor of 100μF has an impedance of 0·16Ω at 10kc/s, comparable with the output impedance of the stabilizer under d.c. conditions.

Conclusion

A stabilizer handling moderate powers and using crystal devices manufactured in this country, has been built and operated satisfactorily. Improvements are envisaged by the use of higher powered transistors when they become available. It is of interest to note that the Zener diode has a comparatively low impedance (1·5Ω) for a current range of 100mA, and these diodes would provide adequate stabilization in many cases if the impedance and current handling capacities could be scaled appropriately. It is clear from the circuit of the stabilizer which has been developed, that it is desirable for the output voltage to be as near as possible to the Zener voltage.

Thus it is preferable to have a range of reference diodes which possess different Zener voltages. At the present time effort appears to be concentrated on developing a Zener diode in the region of 4 to 6V which can operate satisfactorily at room temperatures for currents up to 0·5A. These diodes may of course be connected in series for operation at higher voltages.

Acknowledgments

The author wishes to acknowledge the help and advice given by the Staff and Research Students in the Electrical Engineering Department of Manchester University.

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Dictation-by-Telephone

The latest dictation-by-telephone system has been installed in the new headquarters in Baker Street, of The Metal Box Co. Ltd. Called the Agavox Teledictation System, it enables big savings to be achieved in the cost of producing letters, memos, etc. The new equipment has been manufactured by M. & L. Haycraft Ltd., and has been virtually tailor-made to suit their requirements following a careful preliminary investigation and research. The initial installation—which is now almost completed—consists of between 100 and 150 telephones on three floors linked to 15 Agavox dictating machines in the typing department.

The dictating machine employed in the system is the Agavox, which uses a magnetic disk as the recording medium. Each dictator has complete remote control over the machine by operating a number of push-buttons. Each Agavox is mounted on an automatic control unit. The latter translates these push-button "orders" into operations on the machine, such as dictation, playback, backspace, etc.

The dictator uses a selector switch to "find" a free dictating machine and he can then proceed with his dictation, hearing any part of it at any time, making any corrections necessary, erasing mistakes and so on. Whenever he reaches the end of a letter he presses an "end of letter" button which automatically punches a pin hole in the paper index strip on the machine. A punch hole will be punched in another column if he presses the correction button. Finally, when he puts the receiver down, final end of dictation pin hole is punched in the strip automatically.

Each recording disk on the machine takes 12 minutes' dictation and when the eleventh minute is reached a warning "tick" tells the dictator that time is nearly up and at the end of the twelfth minute he receives a different tone. If he wishes, he can contact the typing department supervisor by pressing a button and talk to her over his telephone. (The supervisor can, of course, also call the dictator).

In the typing department, panels on the front of the automatic control units have red and green lights which tell the supervisor which machines are free, which are being used, if a dictator is calling her, and which machines have taken dictation and require re-setting.

LETTERS TO THE EDITOR

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

L-Type Iterated Networks

DEAR SIR,—For calculating the voltage-transfer ratio of L-type iterated networks, the present writer has derived a simple and purely arithmetic procedure, published elsewhere[1]. Some recent articles e.g. Pease[2], deal with various methods for approximate calculation of the matrix for the n^{th} power of a 2×2 matrix. Since, however, they employ transcendental functions, they are cumbersome and of limited validity.

The author's method may be supplemented to yield an exact and purely algebraic procedure in which all the elements of the matrix are polynomials, the coefficients of which are the result of an arithmetic tabular operation. The resulting polynomials are exact values and after substitution of frequency values, it is possible to neglect all terms which are sufficiently small.

Iterated ladder networks with L-type members of longitudinal impedance Z and transverse admittance Y have the cascade matrix:

$$C = \begin{Vmatrix} 1+YZ & Y \\ Y & 1 \end{Vmatrix}^{n}$$

Calculating the n^{th} power of this matrix is easily possible using a scheme in which the elements of each successive line result from successive addition of the elements of the previous line. The numbers above the table correspond to the power of the matrix; the coefficients under b are those of the element C_{11}, those under a of C_{12} and C_{21} and those in column b for the power n-1 are those of the element C_{22}.

0		1		2		3		4		5		6		7	
–	*b*	*a*	*b*	*a*	*b*	*a*	*b*	*a*	*b*	*a*	*b*	*a*	*b*	*a*	*b*
	1		1		1		1		1		1		1		1
		1		2		3		4		5		6		7	
			1		3		6		10		15		21		28
				1		4		10		20		35		56	
					1		5		15		35		70		126
						1		6		21		56		126	
							1		7		28		84		210
								1		8		36		120	
									1		9		45		165
										1		10		55	
											1		11		66
												1		12	
													1		13
														1	

The iterative matrix has generally the form e.g.

$$C = \begin{Vmatrix} 1+YZ & Z \\ Y & 1 \end{Vmatrix}^{5} = \begin{Vmatrix} C_{11}{}^{(5)} & C_{12}{}^{(5)} \\ C_{21}{}^{(5)} & C_{22}{}^{(5)} \end{Vmatrix}$$

The individual elements of which are

$$C_{11}{}^{(5)} = 1+15ZY+35Z^2Y^2+28Z^3Y^3+9Z^4Y^4+Z^5Y^5$$

The physical meaning of this element is the no-load voltage transfer ratio in the reverse direction.

Element C_{22} has the same coefficients as the element $C_{11}{}^{(4)}$, i.e.

$$C_{22}{}^{(5)} = 1+10ZY+15Z^2Y^2+7Z^3Y^3+Z^4Y^4$$

The physical meaning of this element is the short-circuit current transfer ratio in the reverse direction.

Element C_{12} is a polynomial with the coefficients given in column a for the n^{th} power, multiplied by Z, i.e.

$$C_{12}{}^{(5)} = Z(5+20ZY+21Z^2Y^2+8Z^3Y^3+Z^4Y^4)$$

and finally

$$C_{21}{}^{(5)} = Y(5+20ZY+21Z^2Y^2+8Z^3Y^3+Z^4Y^4)$$

With this scheme it is possible to calculate the matrixes without loss of the time and the danger of error inherent in drawn out complicated calculations.

Yours faithfully,

JINDRICH FOREJT,
Faculty of Radiotechnics,
Podebrady,
Czechoslovakia.

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Thermistor Multivibrator Circuits

DEAR SIR,—In the correspondence columns of your May issue, Mr. Armstrong suggested an improvement in transistor multivibrator circuits designed to reduce the integrating effect of the cross-coupling capacitors. He does so by inserting large series resistors in the base circuits. While producing some improvement in the back (integrated) edge of the waveform, his circuit is bound to slow down the front edge very considerably. In most cases the input capacitance and the finite frequency response of the transistor have a very significant effect upon the waveforms and in his circuit the loop gain is greatly reduced. In mentioning the frequency limitations of his circuit he seems to be aware of its disadvantages.

A way of obtaining fast back edge without significantly affecting the front edge of the waveform and the maximum repetition frequency is shown in the circuit in Fig. 1 (U.K. Patent Application No. 27787/55). If X_1 goes into the cut-off state the diode MR_1 isolates its collector from the integrating capacitor C_1, resulting in a fast edge. The rise-time at the collector of X_1 is slowed down by 10 to 20 per cent due to the increased

Fig. 1 (below) Multivibrator with collector-isolating diodes.

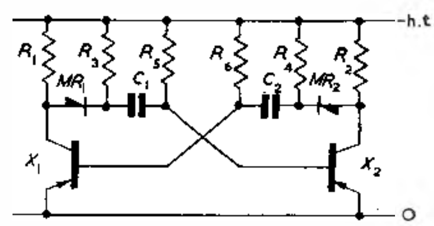


Fig. 2 (above) Collector voltage waveform.

source impedance, viz. R_3 $(=2R_1)$ instead of R_1. If it is desired to take the output from one collector only this deterioration can be minimized by inserting the isolating diode at the output collector only. The result is the waveform shown in Fig. 2.

Yours faithfully,

F. ROZNER,
Ferguson Radio Corporation Ltd.

The correspondent replies:

DEAR SIR,—I agree with Mr. Rozner that the series resistors in the feedback loop are not suitable when fast leading edges are desired in the waveforms, and my comment on frequency limitations was intended to cover this point.

On the other hand, there are some applications for which the waveforms need not be especially fast, but for which they should be as nearly rectangular as possible. It was with such uses in mind that the addition of series resistors was suggested as a simple and economical improvement.

Yours faithfully,

H. L. ARMSTRONG,
Pacific Semiconductors Inc.,
Culver City,
California.

BOOK REVIEWS

Voltage Stabilized Supplies

By F. A. Benson. 370 pp. 249 figs. Demy 8vo. Macdonald & Co. Ltd. 1957. Price 50s.

UP to the end of the war it was unusual to find a piece of electronic equipment that incorporated a hard-valve voltage stabilizer; to-day such stabilizers are in use on equipment of all kinds and a considerable literature has grown up. In this book Dr. Benson reviews the various types of stabilizer, and illustrates his comments with circuit diagrams and performance data taken from papers published in a wide range of journals. The volume contains over a thousand references to original work and for this reason alone it could hardly fail to be of value to engineers concerned with the design of electronic circuits.

The first two chapters deal with simple stabilizers using one or more glow-discharge tubes in conjunction with linear resistances. Worked examples are given to show how these stabilizers are designed and what performance can be expected. The usual difficulty with glow-discharge stabilizers is failure to strike but the need for generous design to give reliable striking is not clearly stated in the text. Chapters III and IV, which account for nearly one third of the book, deal with the characteristics of glow-discharge tubes. This is a field in which Dr. Benson has an especial interest and much of the data refers to work done in his laboratory at Sheffield. Unfortunately there is a fairly wide variation in characteristics between individual tubes of the same type, so that these chapters serve more as a warning not to expect too much from a glow-tube than as a firm basis for design. Furthermore it is possible that the silicon junction diode, running in the "Zener" region, will supersede the glow-tube as a voltage reference device.

Chapter V discusses corona tubes as high-voltage stabilizers for the microampere range, and gives some interesting circuits using cold-cathode triodes for similar applications at medium voltage. Stabilizers of this type do not require heater supplies and are used in portable survey instruments for measuring radioactivity. The conventional degenerative stabilizer widely used for h.t. supplies on amplifiers and pulse equipment forms the subject of Chapter VI. Several practical circuits are given and the use of correction from the unstabilized side to give improved performance is explained. Most of the circuits given use a single pentode shunt amplifier and it would have been useful if this chapter had been expanded to include more information on stabilizers using shunt amplifiers of considerably higher gain. Such stabilizers have better output impedance characteristics and avoid the need for correction from the unstabilized side. Chapter VIII deals with stabilizers suitable for running microwave oscillators; here the treatment is very full and five examples are analysed. The effect of heater voltage variations is studied in considerable detail (page 186-190) but no mention is made of the standard ways in which the effects of such variations may be conquered. Chapters VIII and IX deal with the properties of battery reference elements and of devices working on the magnetic saturation principle.

The final chapter, which runs to over 80 pages, deals with miscellaneous circuits not covered in the earlier chapters. This is undoubtedly a most useful part of the text as it gives notes on a wide variety of published circuits, many of which are not widely known. Information of this type is particularly valuable to the experienced engineer as it is usually possible to do the detailed design once a suitable type of stabilizer circuit has been chosen.

The references are complete with titles and there is an adequate index; however, the references are so numerous that it would have been of further assistance to the reader if they had been classified on the basis of subject matter.

This is a most welcome source book and in view of the importance of its subject matter it will be widely used.

V. H. ATTREE.

Introduction to Printed Circuits

By Robert L. Swiggett. 101 pp. 83 figs. Demy 8vo. John F. Rider Publisher Inc., New York, Chapman & Hall Ltd, London. 1957. Price 21s.

THIS is an excellent profusely illustrated brief introduction into the many methods of making printed circuits and the basic manufacturing set-ups for them. It is kept entirely to the practical side of the present stage of usage of printed circuits, and Mr. Robert L. Swiggett, the author, who is the engineering head of Photocircuits Corporation, New Jersey, is surely the American technical authority with the widest experience in this young industry.

The book starts with a description of the principle of ceramic based circuits as marketed by Centralab and produced by Project "Tinkertoy". It then goes on to describe under the heading of "Etched Circuits" the foil method which is the main process in use in the United States, and it illustrates the basic machines and steps in the production of the raw material and the circuits including their various finishes, such as plating through holes and plating of etch resists.

It briefly describes special punching machines and what is called printed components (switches, capacitors, coils). The plating method as distinct from the etching method of pattern production, and the adhesion problem encountered with this method are lined out. It might perhaps have been desirable to show on this example not only its limitation to—even for American conditions—large volumes, but also its limitation to wiring boards and similar sorts of application.

The fourth chapter of the book reviews some obviously "odd" processes: stamped circuits, the pressed powder method, metal spraying and a punching process. A more promising transfer process is also mentioned.

Chapters V and VI of the book briefly describe the general particulars of components, automatic assembly systems including dipsoldering. The last chapter gives instructions for servicing printed circuits.

The usefulness of an index for such a brief introductory book is perhaps not obvious, but the many photographs and particularly the multitude of clear and self-explanatory diagrams which have been made available by courtesy of Photocircuits Corporation are particularly welcome.

PAUL EISLER

Progress in Nuclear Physics Volume 5

Edited by O. R. Frisch. 300 pp. Demy 8vo. Pergamon Press, New York and London. 1957. Price 80s.

IT may at first sight seem paradoxical that just when the practical uses and abuses of nuclear energy are becoming firmly established, the theoretical side of nuclear physics seems to be running into its stickiest patch yet. Hardly since the days of Ptolemaic epicycles have we seen such an *ad hoc* multiplication of basic entities and properties, and nuclear theorists are the first to admit their dissatisfaction with the situation today.

But the paradox is only apparent, as the present volume may help to show. Men learned to burn coal centuries before the physical chemistry of hydrocarbons was unravelled, and we may hope that the exploitation of nuclear energy will similarly progress despite our perplexity as to the best theoretical model.

As may be expected, some of the contents of this fifth Progress Report are barely intelligible, or of little interest, to the non-specialist. J. M. C. Scott's article on the Radius of a Nucleus and B. W. Ridley's on the Neutron, however, in their less technical sections, offer illuminating glimpses into the ferment of discovery and speculation which characterizes the field.

Papers on experimental techniques for the measurement of reaction energies (W. W. Buechner) and the study of the inelastic scattering of fast neutrons (Joan M. Freeman) provide a useful background to the article which will be of chief interest to electronic engineers, on Electronic Techniques for the Nuclear Physicist, by K. Kandiah and G. B. C. Chaplin. This is notable for a long section of 20 pages on the transistor as a circuit element, which ends with a brief but balanced review of likely developments, as of 1954.

One challenging observation is that "the demanded degree of reliability" (in nuclear instrumentation) "is higher than that achieved in electronic equipment in any field at present" (page 90). The ingenuity revealed by the circuits here reviewed shows that electronic engineers have not at any rate been idle in face of these new needs. It will be rewarding reading for anyone interested in the limits to which electronic techniques can be pushed.

The production is of the usual high standard, and apart from one misprint (on page 98) and an occasion lapse from grammatical usage of the kind which is now all too familiar in technical writing, no blemishes distracted the reviewer.

D. M. MacKay

Präzisionsmessungen von Kapazitäten, Induktivitäten und Zeit-Konstanten (Precision Measurements of Capacitances, Inductances and Time Constants).

By Dr. Erich Messerschmidt. 166 pp. 84 figs. Small 8vo. Friedrich Vieweg & Sohn, Braunschweig. 1956. Price DM.11.80.

THIS slender volume is the second edition of a monograph published in 1940. It deals with capacitances, dielectric losses and permittivities only while the measurement of inductances will be treated in a subsequent booklet. It aims at describing well tried methods of precision measurement with all experimental details and thus is suitable also for those physicists and engineers which are no specialists in this field but are occasionally called upon to perform such measurements. For them the selection of a suitable method is facilitated and their attention is drawn to the numerous sources of error.

A brief theoretical survey deals with capacitances and partial capacitances, permittivities, the units used, the calculation of capacitors and the losses. This chapter contains a detailed table of permittivities and loss factors of a large number of solid, liquid and gaseous materials. The second chapter describes the various standards of capacity in great detail and illustrates them in halftone pictures and line drawings. In the third chapter the precision methods for measuring capacitances are discussed. First the direct methods are described and then the indirect methods including a great variety of bridge methods. Special attention is paid to high frequency measurements. The measurement of permittivities of solids, liquids and gases is discussed.

A special feature of the book is a very extensive bibliography of nearly 20 pages covering the literature up to 1954. A name and a subject matter index conclude this useful little monograph. Unfortunately the pages given in the name index are not always quite correct but differ by one or two pages.

R. Neumann

Variable Capacitors and Trimmers

By G. W. A. Dummer. 169 pp. 55 figs. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1957. Price 32s. 6d.

IN conformity with the three books on components recently published by the author this further volume comes up to the fullest expectations.

Its 169 pages embrace ten chapters containing many facets of the theory, design and manufacture of variable capacitors and trimmers. As in the previous volumes, Service type components have been chosen as representatives of high standard.

Commencing with general specifications and a brief history the chapters progress through measurements on variable capacitors, general purpose, precision and transmitter capacitors to trimmers and special capacitors such as phase-shifters and butterfly types.

Faults which may occur are dealt with, also there is a short chapter on future developments in the light of current trends.

The book concludes with a very comprehensive bibliography of some 184 references followed by 36 pages of comparison charts of variable capacitor characteristics, both electrical and mechanical, that should be of considerable worth.

The work demonstrates how necessary it has been for someone to correlate all the miscellany of available information on a wide subject and the author seems to have accomplished this very thoroughly.

The chapter on measurement deals with methods using frequencies from low values to those in excess of 3 000Mc/s. Descriptions of measurement processes are somewhat meagre as the precision measurement of capacitance over a wide frequency range is a complex subject but references to more detailed information can be obtained from the bibliography.

On the other hand, some fairly detailed tables are included such as the temperature and humidity charts in Chapter I and the properties of dielectric materials in Chapter II.

The volume is essentially a practical one dealing as it does in laws of operation, methods of manufacture and construction.

Innumerable photographs and pictorial representations abound throughout. In fact, there is scarcely a page that does not contain an illustration of some kind; all of considerable interest to a casual reader but much of it of prime importance to the design engineer.

R. H. Mapplebeck

Basic Mathematics for Radio and Electronics

By F. M. Colebrook. 359 pp. 90 figs. 3rd Edition. Demy 8vo. Iliffe and Sons Ltd. 1957. Price 17s. 6d.

THIS book was previously published under the title *Basic Mathematics for Radio Students*. The new edition is wider in scope and includes much new material.

ELECTRONIC EQUIPMENT

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

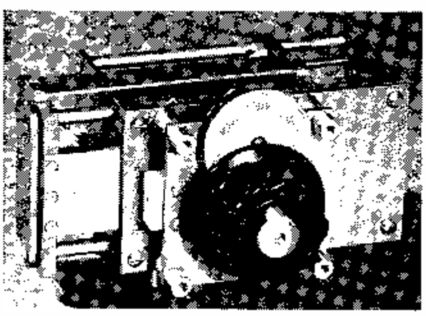
TURRET ATTENUATOR

(Illustrated below)

Advance Components Ltd, Roebuck Road, Hainault, Essex

THE type A63 is a completely coaxial attenuator unit having six coaxially-constructed resistive pads which are selected by rotating the control knob. The attenuation of the pad selected is indicated by a figure which appears on a dial geared to the control. The selection is effected by a mechanical linkage system which unplugs the coaxial connexions to a pad, rotates the turret to the desired position and plugs into the pad selected with only one rotary movement of the control knob.

The attenuator may be fitted with any combination of pads from 0 to 50dB in 10dB steps.



Input and output impedances are 75Ω and the attenuator is designed to accept standard Plessey Minor coaxial fitting.

The attenuator is manufactured to very close tolerances and each unit is subjected to vigorous tests at all stages of manufacture. It is intended for use in high class equipment where absolute reliability and rapid selection is required. The attenuator is capable of handling continuously up to one watt of sine-wave power. In sharply pulsed conditions its rating is limited by the peak voltage of the signal. The limit of this voltage is of the order of 100V for pulses of a few microseconds duration.

A maximum attenuation of 50dB is recommended at 3 000Mc/s, but this may be extended at lower frequencies. The temperature coefficient varies with pads used and is of the order of 0·205dB. Accurate figures for the pads used and temperature range required will be supplied upon request.

FREQUENCY AND TIME MEASURING EQUIPMENT

(Illustrated bottom right)

Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey

VENNER Electronics Ltd are now in production with a transistorized frequency and time measuring equipment incorporating plug-in units. The equipment is extremely versatile and is capable of measuring any frequency in the range 10c/s to 50kc/s, the period of any waveform from 0·00001c/s to 10kc/s, and the time interval between two consecutive pulses from the same time range. Digital presentation is employed, the indicator for each decade being a meter calibrated from 0 to 9.

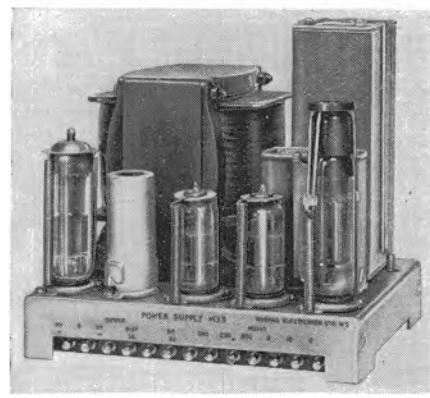
When set to measure frequency, any of three gating times may be selected, 0·1sec, 1sec or 10sec. The count of the unknown frequency will then be carried out for any of the selected periods. In this way, the frequency to one decimal place may be determined. Six standard output frequencies are available and any of the six frequencies may be measured by depressing the test switch. An automatic timer is fitted so that repetitive measurements may, if desired, be made without the equipment being operated by hand. The display time is variable from 0·5sec to 5sec.

When set to period measurement the internal 10kc/s crystal oscillator is counted for the duration of one input cycle of the unknown frequency. In this way, an extremely high order of accuracy can be obtained for frequencies of from 0·00001c/s to 100c/s. Once again, the display timer may be employed, enabling repeat counts to be made. The test switch is still operative, permitting the internal reference frequencies to be injected.

With the equipment set to pulse interval timing the two inputs provided become operative and a pulse fed to the start socket opens an internal gate permitting the feeding of any of the six internal frequencies to the counting stages. Thus, if 1c/s was selected, the time-base would be in units of 1sec and the total indicated on the counting decades before a pulse is fed into the stop socket would read directly in seconds. If, however, the 1kc/s output was selected on the time unit output frequency control, then the units of the indicated count would be in milliseconds. This permits a very wide range of timing to be carried out with optimum accuracy obtained at all times. Once more the display timer may be made to operate so as to hold the indicated elapsed time for an adjustable period between 0·5sec and 5sec, at the end of which it resets all decades to zero awaiting the receipt of another start pulse.

The contact timing position on the function selector enables ten different types of timing to be carried out from virtually any combination of contacts. It will, for example, measure the closed or open time of a single pair of contacts or the interval between one pair of contacts closing or opening and another pair of contacts opening or closing. Once more, any of the six internal time-bases may be selected.

The techniques used in this equipment are digital throughout, both as regards the counting and the divide-down stages.



MINIATURE REGULATED POWER SUPPLY UNITS

(Illustrated above)

Roband Electronics Ltd., 33 Mountgrove Road, Highbury, London, N.5.

THESE units are constructed on open chassis and may be used on the bench and also in enclosed instruments. Each has a highly stabilized fixed d.c. output voltage, either rail of which may be earthed, and the output current may be varied from zero to the minimum specified value without noticeable voltage deviations. D.C. stability is maintained for a mains variation of ±10 per cent. Two isolated 6·3V heater supplies are also provided.

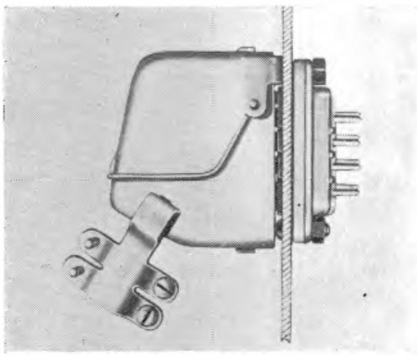
A particular feature is the omission of the conventional iron core choke in the rectifier section. Apart from the obvious saving in weight and size there are further advantages in that instantaneous load changes do not produce voltage shock excitations, and also the output current can be reduced to zero without instability. The design is such that, notwithstanding the omission of the choke, the average unit has a ripple content of well under 1mV.

A preset control enables the source impedance to be varied smoothly from positive to negative values; each unit is adjusted to zero before leaving the factory. If, however, it is desired to operate the supply in series with a milliameter it is possible to adjust the control such that the impedance, including the meter, is zero.

Interservices approved paper capacitors are used throughout and units will operate at elevated temperatures and after a long shelf life. Conservately rated transformers, valve operation well within limits and the use of high stability carbon resistors in preference to wirewound types ensure prolonged trouble-free operation.

The type M33 is a typical unit and has the following characteristics.

Mains input ..	200/250V.a.c. 40-60c/s
D.C. output voltage	±250V.
D.C. output current	0 to 50mA.
Effective resistance..	Less than 0.25Ω
Ripple	Less than 1mV.
Voltage stabilization	Less than 0.02 per cent
A.C. output ..	6.3V at 2A. 6.3V at 1A.
Size	6in × 5½in × 6in high.



MULTI-CONTACT PLUGS AND SOCKETS

(Illustrated above)

Carr Fastener Co. Ltd Distributed by: The Benjamin Electric Co. Ltd, Tottenham, London, N.17

CINCH 'J' type plugs and sockets are available in 4, 8, 12, and 20 way versions. They include contacts of a special design having twelve resilient fingers which engage with each flat plug pin to provide low contact resistance, high current carrying capacity and minimum insertion and withdrawal forces.

The mouldings are of Mikacin, a thermo-plastic material, which is non-tracking, has low moisture absorption, low loss, a temperature range of 40°C to 100°C and high insulation resistance.

A spring steel clip ensures positive retention of the plug when mated with the socket and affords efficient earthing of the cover screen. The overall dimensions of these components are comparatively small.

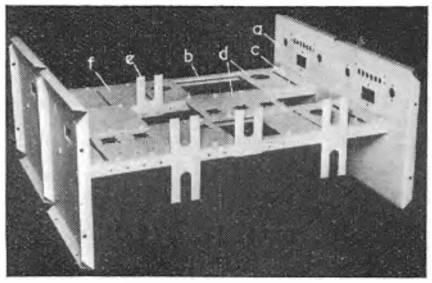
STANDARD CHASSIS

(Illustrated above right)

Cowell Developments, 67 Long Drive, East Acton, London, W.3

THIS versatile, non-expendable metalwork system is designed to replace the various temporary chassis, panels, etc., required during circuit development and experimental work, or for the construction of 'one-off' test gear, educational apparatus, etc.

The various parts, of multi-purpose



design, enable the mounting of a wide range of components. The design is mechanically robust (all parts mild steel, cadmium plates), permitting immediate modification and extension of any particular arrangement during use. Close approach to a desired layout is therefore possible.

The system is available in unit kit form, a unit measuring 7in × 7in × 16in. Each kit enables the construction of an appropriate chassis for circuits containing up to eight valves, etc., and any number of units may be joined together for large equipments. Units, or groups of units, may be fixed directly to standard 19in rack panels.

The end plates enable mounting of power supply connectors, coaxial or jack sockets, toggle switches and terminal strip. The runners are drilled and tapped 4 B.A. (holes spaced ¾in) enabling the mounting of long tag strips with either ⅜in or ¼in spaced tags. The two kinds of valveholder plates afford mounting for the majority of valve types. The potentiometer plates provide fixing for potentiometers, toggle and rotary switches, jacks, etc. The blank plates provided may be cut to accept special components.

Illustrated is a typical assembly (shown inverted) the components shown being

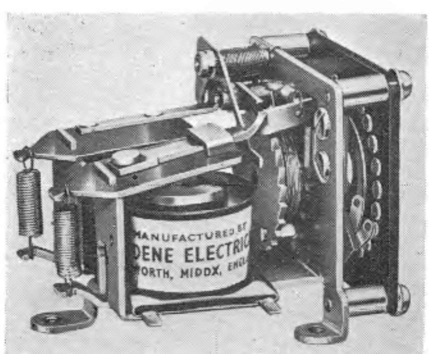
(a) End Plate
(b) Runner
(c) Bracket
(d) Valveholder plate
(e) Potentiometer bracket
(f) Blank plate

ROTARY STEPPING RELAY

(Illustrated below)

D. Robinson & Co., 58 Oakes Avenue, Worcester Park, Surrey

THE 'Rodene" stepping relay is intended for use as a selector, counting or memory relay in such applications as the automatic control of apparatus and production processes, this unit has a



rotary switch with heavy stud type fine silver stationary contacts and silver-graphite moving contacts.

It can be either 20 way single pole, or 10 way double pole, and is stepped forward one position every time a pulse is fed to the coil shown in the front of the photograph.

A momentary impulse to a second coil lifts a pawl and allows a spring to reset the switch to the zero position. This is assured by a latch which holds the pawl out until the next forward pulse.

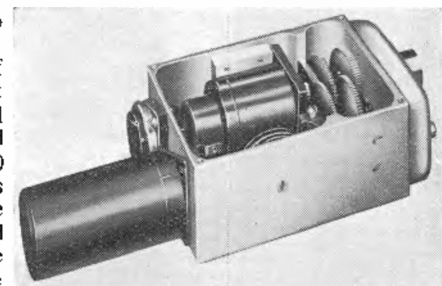
Important features are that the unit can be operated direct from a.c., and that the consumption is only 45VA on a.c., or 15W on d.c. The maximum speed is 40 steps per second.

SERVO UNITS

(Illustrated below)

Harvey Electronics Ltd, 273 Farnborough Road, Farnborough, Hampshire

THESE servo units have been specifically designed to meet the increasing demand for an extremely light yet accu-



rate power amplifying device which will provide a displacement output precisely related to an input voltage.

Since the original application of the servo units was the remote radio control of experimental pilotless aircraft, a rapid response was also essential, and the transit time of standard units is less than one second, with maximum angular rates approaching 100°/sec.

Each unit consists of a machined cast magnesium box containing the driving mechanism—a small d.c. electric motor operating a stainless steel shaft through a train of specially designed spur gears. The motor is controlled by a miniature relay with special long life contacts which plugs into one end of the box and the position of the output shaft is detected by a precision wire-wound potentiometer mounted on the other end of the box. Power and signal information are fed in through a connector mounted alongside the relay.

All components are thus available for servicing, and provision is made for adjustment of the gears to eliminate backlash.

Brief details of the performance are as follows:

	TYPE 0	TYPE 1
Stalled torque	1 lb-ft	5 lb-ft
Maximum speed, no load	60°/sec.	80°/sec.
Input	±6V	±24V
Repeatability	0.1% full scale	0.1% full scale
Deflexion for full load (90% stalled torque)	2% full scale	2% full scale
Deflexion	±28°	±28°

Short News Items

Higher Technology Courses at Southall Technical College, with starting dates and duration in weeks, are as follows: Monday, 30 September, Radio Telemetry (12); Tuesday, 1 October, Digital Computers (12); Servomechanisms Theory (24), Microwave Theory (24); Wednesday, 2 October, Transistors (12), Experimental Servomechanisms (12); Monday, 7 October, Advanced Mathematics (24); Wednesday, 9 October, Analogue Computers (10); Thursday, 10 October, Pulse Techinique (18), Industrial Electronics (18); Monday, 13 January, 1958, Sound Recording (10), Reproduction (12); Wednesday, 15 January, Colour Television (10), Cathode Ray Tubes (12). Further details may be obtained from the Head of the Electrical Engineering Department, Technical College, Beaconsfield Road, Southall, Middlesex.

Borough Polytechnic have arranged a course of experiments in the Pulse Techniques Laboratory, under specialist supervision, on Monday afternoons or Thursday evenings, beginning 28 October and 31 October, respectively. The course will extend over twelve weeks. Applicants should have theoretical knowledge to a standard reached in the Polytechnic Lecture Course on the subject. The fee is £1. A course of twenty-two lectures on the Fundamental Principles of Pulse Techniques will be given by specialist lecturers, beginning on Monday, 7 October. These lectures will begin at 7 p.m. on Monday evenings. The fee for this course is £2 10s. Enrolment forms may be obtained from Borough Polytechnic, Borough Road, London, S.E.1.

The Wembley Evening Institute announces that classes will again be held, as last year, to prepare candidates for the City and Guilds Radio Amateurs' Examination. There will also be morse practice classes. The classes will be held on Mondays and Thursdays, Morse 7-8 p.m., Radio Theory 8-10 p.m. Enrolment will be at Copland School, High Road, Wembley, on Monday 16 September to Thursday 19 September between 7 and 9 p.m. and the classes will commence during the week beginning 23 September.

The Carr Fastener Co. Ltd. of Stapleford, Nottingham, have opened two new factories at Worksop, and at Sutton-in-Ashfield. The latter is a moulding plant which has been built to increase the production of this unit of the company by about 250 per cent. In addition to press fasteners, a large variety of spring clips and other automobile parts, the Carr Fastener organization turns out well over three million radio and electronic assemblies each week.

The Independent Television Authority has placed contracts worth approximately £160 000 with Marconi's Wireless Telegraph Co Ltd for the transmitting equipment for four new stations. One set of equipment will be installed in the Chillerton Down (Isle of Wight) transmitting station of the ITA, which is planned to come into service next summer. The other three have been ordered as part of the long-term plans of the Authority, and it is likely that they will be installed in stations to be opened in North-East England, the Cumberland area and Northern Ireland. Each station will be equipped with two 4kW vision transmitters, two 1kW sound transmitters, combining units, programme input equipment and a comprehensive amount of ancillary equipment. At two of the stations the Marconi high-power parallel-operated system, which has proved a success at the London station, will be used, while at the remaining two stations it is provisionally planned by the Authority to operate one vision-and-sound pair as "Main" with the other pair as "Standby". The 4kW vision transmitter is of a new design.

The Wayne Kerr Laboratories have recently entered into agreement with the Robertshaw-Fulton Controls Co of Pennsylvania whose Instrument Division in Philadelphia are primarily concerned with the manufacture of industrial process control instruments. The agreement between the two companies is in two parts. In the first Robertshaw-Fulton will be responsible for the promotion and distribution of Wayne Kerr products in the U.S.A. but they will continue to use the present agents. In the second part of the agreement Wayne Kerr will license Robertshaw-Fulton to use their patented measurement techniques in the field of industrial process control and the two companies have agreed upon a joint development programme to achieve this object. The resulting products will be manufactured by Robertshaw-Fulton in the United States and by Wayne Kerr in England. An intitial order exceeding $100 000 in value has been placed by the Robertshaw-Fulton Co to launch the agreement.

The first of a series of Decca Windfinding Radars ordered for the Argentine National Meteorological Service has recently been delivered at Buenos Aires. These Decca radars are specifically designed for upper air windfinding and are the only equipments of their type available in the world today. They will be used to form a network of upper air windfinding stations to provide data from heights up to 100 000ft for the forecasting and research staffs of the Argentine National Meteorological Service.

Ferranti Ltd have announced that a new Pegasus electronic digital computer was recently handed over officially to the Northampton College of Advanced Technology by the Managing Director of the National Research Development Corporation, Lord Halsbury, to celebrate the College's higher technological educational status. Previously the College was known as the Northampton Polytechnic.

At the Het Atoom Exhibition, which is being held at Schiphol Airport in Holland until 16 September, Pye Ltd are exhibiting a range of equipment of nuclear applications, including slave manipulators, industrial television and an underwater television camera operating in a "swimming pool" reactor.

Lord Brabazon of Tara has been elected President of the Radio Industry Council in succession to Sir Edward Appleton.

Mr. G. B. Campbell has been appointed secretary of the Radio Industry Council in succession to Mr. R. P. Browne, who has retired through ill health.

Mr. D. C. Birkinshaw, Chairman of the Television Society Council, and Mr. T. H. Bridgewater, the Honorary Treasurer, have this year completed twenty-five years in the BBC Television Service. This may fairly establish their claim to be the first television engineers in this country, as they joined the BBC in order to operate the first low definition 30 line television developed by the Baird Company. Mr. Birkinshaw is now Superintendent Engineer Television, and Mr. Bridgewater is Superintendent Engineer of Television Outside Broadcasts.

The Institute of Physics' telephone number is now Belgravia 6111.

Erratum. The penultimate paragraph of the note "A Direct Recording Cathode-ray Tube" on page 373 of the August issue should read ". . . . with 1 000 points in 10cm linear . . . .".