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Commentary

FEW of us can remain unaware of the low state of efficiency into which the British railways have fallen in recent years and while the politicians have been arguing the respective merits of nationalization and denationalization the general public has come to regard the transport system of this country as a heavy liability rather than an asset.

That the railways in particular have run down no one denies, and prominent among the causes are the severe usage to which the railways were subject during the war and the subsequent lack of capital investment which has prevented the replacing of its worn-out and obsolete equipment.

However, the ills afflicting the railways are soon to end for the British Transport Commission "convinced that an efficient and modernized railway system is essential to the economy of the country" have recently laid before the Minister of Transport a plan by which it hopes "to produce a thoroughly modern system fully able to meet both the current traffic requirements and those of the foreseeable future".

The plan involves the outlay of no less than £1 200 million and during the next fifteen years will produce a complete transformation of the railway system as we now know it. At the end of that period we shall have said farewell to the steam locomotive and welcomed in its place the 100 m.p.h. electric or diesel unit.

At this early stage, full technical information on the proposals is naturally lacking but we are assured that "behind this brief presentation there lies a mass of detailed study" and no doubt we shall learn more as the plans proceed.

Of particular interest is the section dealing with signalling and telecommunications. It is proposed among other things to modernize the existing telegraph and telephone systems and automatic train control equipment at a cost of some £100 million and the Commission promises "to take advantage of all available developments in telecommunication systems".

As far as communications between fixed points of the railways are concerned no serious criticism can be levelled for an adequate and satisfactory system has been built up. It is a fact of course that the railways were among the early pioneers of communications and it is worth recalling that the first inland telegraph system, probably in the world, was installed between Paddington station and Slough as far back of 1838.

But in the field of mobile communication systems it would appear that little or no progress has been made, and the railways have fallen far behind current practice in America where very extensive use has been made of v.h.f. radio for communication purposes between drivers, train crews and despatchers in marshalling yards. Experiments have been carried out by the British railways since the war with v.h.f. mobile equipment with little success for it was obvious that the steam locomotives on which the trials were carried out were far from ideal for the purpose. The engine cabs were dirty, and the noise made intelligibility difficult. Power supplies were not easily generated and suitable aerial arrays could not be erected.

But these are disadvantages associated only with steam locomotives and no doubt these points will be taken into account in the design of the future diesel and electric units.

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In its annual report for the year ending June 30th 1954, which has just been published, the National Research Development Corporation states that the electronic digital computor is still the largest single project undertaken by them and it is their intention to give their full support to its further development for some time to come.

Of the total sum of nearly £1 million borrowed by the Corporation up to date, over £325 000 is accounted for by its work on computors, and further expenditure is envisaged.

This is encouraging news for although expenditure in the U.S.A. on computors is believed to be many times greater, the Corporation is of the opinion that this country is in no way falling behind in the development of this intriguing machine.

To date six large computors of the type installed at Manchester University have been built or are being built for the Corporation's account and interest is now being centred on a much less costly machine, namely the 401 Mark 1. This computor* was exhibited for the first time at the Physical Society Exhibition in 1953 and after completion of its trials at the Mathematical Laboratory of Cambridge University, it was transferred to the Rothamsted Experimental Station where it is now in regular use for calculations relating to agricultural statistics.

The Corporation has in addition approved the design of a further type of computor to be known as the FCP1. This "packaged" computor, brief details of which are given elsewhere in this issue, embodies many of the successful features of the prototype 401 together with improvements concerned as a result of programming experience.

* *Electronic Engineering*, 25, 201 *and* 213 (1953).

Some Aspects of V.H.F. Sound Broadcasting and F.M. Broadcast Stations

(With Reference to the Performance of the BBC's Wrotham V.H.F. Station)

(Part 1)

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In this article the relative merits of a.m., a.m.l. and f.m. for a v.h.f. broadcasting service in this country are considered and the results of field tests are given, from which it is concluded that f.m. is the best choice. The Wrotham high power experiment is described in detail. The design of f.m. transmitters is considered with special reference to f.m. modulators and monitoring. V.H.F. aerial and transmission line systems and the parallel operation of f.m. transmitters are dealt with. Finally the unattended operation of transmitters is discussed.

THE use of very high frequency broadcasting for augmenting the number of channels available for sound broadcasting in Europe has been contemplated for many years. Even before World War 2 the number of medium frequency channels available to the BBC for the United Kingdom was proving insufficient for the national distribution of its programme services and is now quite inadequate for the distribution of the present Home, Light and Third Programmes. Interference from the greatly increased number of stations in Europe using the same or an adjacent channel, particularly at night, fading and the need to use synchronized common-frequency working have significantly restricted the useful service areas of the BBC's m.f. stations; serious interference is frequently experienced up to the 15mV contours of some BBC channels during conditions of night-time propagation. Attempts to improve matters by re-allocating operating frequencies and adjusting effective radiated powers on an international basis have not met with much success, and it seems that the situation must deteriorate and national coverage be further impaired. For this reason the BBC deemed it essential to make extensive field trials at v.h.f. and so be in a position to make definite proposals for the provision in this country of a sound broadcasting service using the v.h.f. band of 87·5 to 100Mc/s (Band 2) allocated for the European Region.

The planning of a v.h.f. service involves a knowledge of the mode of propagation of the frequencies concerned[1] which differs appreciably from that at m.f. where earth conductivity and the reflecting properties of the ionosphere have a major influence on the service area of a station and on the ability of stations mutually to interfere with one another[2]. At v.h.f. some of the limitations imposed by these factors are avoided, the primary factors affecting service area being the influence on field strength, under conditions of standard propagation[1,3] of the contours of the terrain between the transmitter and receiver (which produce field strength "shadows" lower than the average field for the neighbourhood) and the local effects produced by screening, absorption and reflexion of the signal by buildings and similar objects.

There are also second order effects produced by non-standard tropospheric reflexion[1] and changes in meteorological conditions which may cause wide temporary fluctuations in field strength and, finally, the annoying "flutter" produced by aircraft local to the receivers[4]. These factors cannot be discussed here, but in practice the proper siting of the transmitting stations[5] can do much to mitigate some of these influences. A v.h.f. service, while not the panacea of all ills, can, nevertheless, alleviate many of the difficulties arising from the present overcrowded situation including the necessity for deliberately restricting the bandwidth receivers to minimize adjacent channel interference. Such a service also has the important general advantage of greater freedom from interference by electrical disturbances, although the regular impulsive type of interference produced by motor-car ignition systems can be objectionable when the signal is weak, and must be combated. It follows that listeners can expect improved programme quality with low background noise, as the viewers of the BBC's v.h.f. Television Service in Band 1 are already aware.

In this country the television service uses conventional amplitude modulation for the sound transmission, and while this is the only practicable system for m.f. broadcasting a choice of systems is possible for v.h.f. broadcasting when the bandwidth of each channel need not be so restricted. It is, however, of first importance to keep the cost of receivers low and on this score the systems of modulation which merit consideration are a.m. (amplitude modulation), a.m.l. (amplitude modulation with a noise limiter incorporated in a wideband receiver) and f.m. (frequency modulation). The various forms of a pulse modulation used for multiple-channel communication systems are not considered to offer sufficient advantage for a broadcast service. With each of the above-mentioned systems the inherent quality of programme reproduction can be sufficiently high, being limited mainly by the response of the audio frequency stages of the receiver and the loudspeaker which are common factors. The most realistic basis of system comparison is the relative extent to which they suppress noise and interference at the receiver, viz. impulsive noise of the type produced by motor-car ignition systems, receiver hiss, atmospherics and the adjacent channel and common channel interference produced by other transmitting stations.

Using a.m. as a basis, it is necessary to specify for f.m. the frequency deviation and the pre-emphasis time-constant to be used, because the overall performance of

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the f.m. system is very dependent on these factors. For a.m.l. the performance is dependent on the radio frequency bandwidth of the receiver, which for comparison purposes should be the same as for the f.m. receiver. Pre-emphasis at the transmitter with de-emphasis at the receiver could be used with a.m. and a.m.l., but it is of real advantage only in f.m. because of the " triangular " noise spectrum of this system. Even so, its use in f.m. requires a reduction in the average level of modulation to avoid over-modulation at the higher frequencies, so that in practice the advantage of pre-emphasis in improving the signal-to-noise ratio is not so great as might at first be expected.

First-hand information about the relative overall performance of a.m., a.m.l. and f.m. at v.h.f. and on the most appropriate transmission constants was obtained from the field trials made by the BBC during 1945[6]. These were carried out with low-power transmissions (about 1kW e.r.p.) on both a.m. and f.m. at frequencies in Bands 1 and 2 and were mainly devoted to propagation tests (field strength versus distance, for both horizontal and vertical polarization) and comparative assessments of f.m. versus a.m. and a.m.l., particularly in regard to signal-to-noise ratio. The optimum transmission characteristics for f.m. were thought to be 75kc/s deviation and 50μsec pre-emphasis, and with regard to noise it was confirmed beyond doubt that, compared with a.m., f.m. gives a considerable reduction of both impulsive noise and receiver hiss. On the same basis a.m.l. gives a reduction in impulsive noise, but not to the same degree as f.m., and no improvement in hiss. The following short notes are considered relevant.

Impulsive Noise

It is difficult to be precise about the relative abilities of the various systems to combat impulsive interference. To achieve the optimum suppression with f.m. careful selection of the time-constants of the receiver circuits which precede the discriminator is necessary. Again, the degree to which impulsive noise is suppressed in f.m. depends upon the ratio of peak interference to peak carrier. Measurements using repetitive uniform impulses as the source of noise show that, provided the peak interference to peak carrier ratio is less than unity, f.m. gives an improvement of about 25dB over a.m. in relative audio signal-to-noise ratio. At ratios greater than unity corresponding to signal levels below the so-called " improvement threshold " the improvement falls away to a minimum value of about 12dB, but thereafter progressively rises again as the ratio of peak impulse to peak carrier increases.

The comparative performance of a.m.l. lies between those of f.m. and a.m. and tends to approximate to that of f.m. for a limited range of peak impulse to peak carrier ratios below unity. Fig. 1 is an experimental result which is also considered to be in agreement with subjective assessments of annoyance produced by regular impulsive interference. When actual motor-cars are used as the source of interference the curves are smoothed out but the general trend is the same, and f.m. still shows marked improvement over both a.m. and a.m.l.

Receiver Hiss

Because of the considerably lower field strengths at which receivers can operate, due to the improved suppression of external interference, receiver hiss becomes a matter of some significance. It can, in practice, be one of the factors which limit the range of a v.h.f. service, and the comparative performance of modulation systems on this score is important. With an f.m. system, using 75kc/s deviation and 50μsec pre-emphasis, the level of receiver hiss is theoretically 26dB lower than with a.m. or a.m.l. provided the input signal level is sufficiently high to operate the limiter. This improvement can be realized in practice by the use of a well-designed receiver.

Expressing this result in terms of the signal field strength required to give a service in which the background hiss is classed as " just perceptible," it has been concluded that a satisfactory service can be obtained with f.m. at a field strength as low as 50μV/m, whereas the corresponding figure for a.m. or a.m.l. is at least 900μV/m. These figures depend to some degree on the operating frequency and in this case apply to Band 2. It may be argued that the advantage of f.m. in suppressing receiver hiss cannot

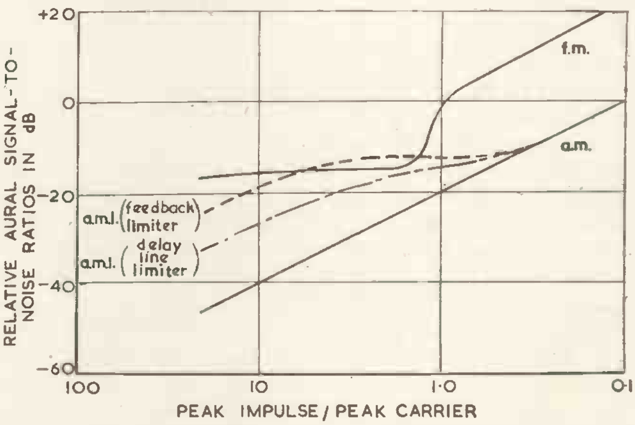

Fig. 1. Impulsive interference: experimental comparison of f.m., a.m. and a.m.l. receivers

usefully be exploited in practice because in combating motor-car interference the ambient signal strength would need to be appreciably greater than 50μV/m. This is to some extent true, and the BBC proposal, for example, is to provide an average (median) field strength of 250μV/m at 30ft above ground level for second-class service. Nevertheless, considerable point-to-point variations in this field strength can occur, while listeners will undoubtedly attempt to receive the service with indoor aerials, for which the field will be considerably reduced. If a relatively wide audio frequency bandwidth, say 12kc/s, is used, the level of receiver hiss is increased and critical listeners become aware of receiver hiss if the signal-to-noise ratio is less than about 60dB.

Horizontal versus Vertical Polarization

There appears to be little difference between horizontal and vertical polarization as regards field strength itself, but the degree of interference produced by motor-cars appears to be significantly less with horizontal polarization. This is not fully understood and may be dependent, to some degree, on the relative directional properties of the receiving aerials. Comparative tests on Band 2 made in relatively open country show that, in general, horizontal polarization gives, roughly, an 8dB improvement in freedom from motor-car interference, but this does not apply under the confused conditions which can exist in closely built-up areas. The effects of signals reflected from aircraft flying in close proximity to the transmitter or receiver are more pronounced with horizontal polarization[4], but this is not considered to be of sufficient general importance to influence the decision to use horizontal polarization even if suppressors were universally fitted to all motor vehicles.

Co-channel Operation

With the Band 2 frequencies available for the United Kingdom, common-channel working will undoubtedly be necessary if only because certain stations will have to share channels with stations in Western Europe. The probability of interference between common-frequency v.h.f. stations, mainly due to non-standard refraction in the troposphere occasionally favouring long-distance transmission, has been investigated by the BBC, the Post Office

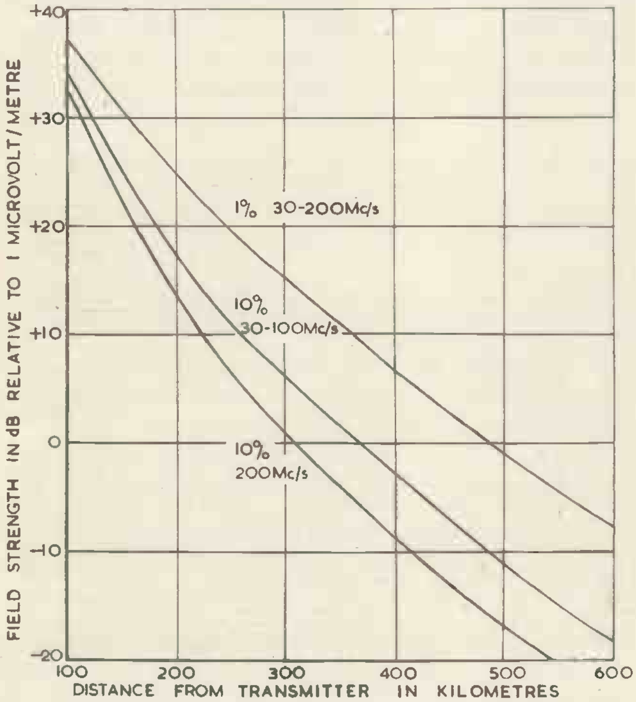

Fig. 2. Curves relating to geographical spacing of v.h.f. broadcasting stations

Estimated tropospheric field strength exceeded for 1 per cent and 10 per cent of the time.

Frequency range 30 to 200Mc/s 1kW radiated from a half wavelength dipole, vertical and horizontal polarization.

Receiving aerial height 10m above ground level, median values with respect to terrain.

and D.S.I.R. over a period of several years. This data has been analysed and applied to the determination of the geographical spacing required between common-frequency v.h.f. transmitters[7]. It is found that if the wanted signal must be, say, 30dB greater than the interfering signal for 90 per cent of the time, the spacings required are several hundred kilometres greater than those normally necessary under standard conditions. The curves of Fig. 2 are relevant, and were prepared from the above data supplemented by data from the U.S.A.

An engineering point here is that while very close synchronism to avoid interference can be maintained between two or more medium frequency stations operating on the same channel, this is not so easily achieved with v.h.f. stations. The co-channel interference may thus be heard as a beat note. A practical way of minimizing this interference is to make the frequency spacing between the centre frequencies of the stations greater than about 12 to 15kc/s so that the beat note is outside the audio frequency range. Interference is then heard only when one or both transmitters are modulated, and takes the form of a rasping noise, the level of which, contrary to a.m., does not depend much on whether the same or different programmes are being transmitted. With 15kc/s separation, the permitted strength of the interfering signal may be about 6dB higher than with a closer spacing, and a ratio of 24dB for the wanted to interfering signal is sufficient for no perceptible interference.

With a.m. this ratio needs to be about 46dB for no perceptible interference when the two signals are modulated with different programmes. For the special case of common programme synchronized co-channel operation, a.m. is better than f.m., a ratio of 15dB being sufficient.

At first sight this practice may appear to contradict f.m. theory, which indicates that the degree of phase modulation, and hence the equivalent f.m. deviation produced at the discriminator of the receiver, increases as the frequency difference between the interfering signals is increased. The point is, however, that although the level of the a.m. output of the discriminator is correspondingly increased its frequency is outside the a.f. response of the receiver, or outside the audible range, so that the interference is not heard. The use of pre-emphasis also leads to an improvement in this connexion because the corresponding de-emphasis in the receiver reduces its response to high interfering beat frequencies.

Adjacent-channel Interference

The channel spacing in Band 2 which has been proposed for stations in this country is 200kc/s for f.m. The degree of interference from a station operating on an adjacent channel depends on the details of the receiver design, and on the actual strength of the wanted signal. With a good commercial receiver and wanted signal of 1mV/m, if the adjacent-channel signal is between 10 to 20dB less no interference is audible. With signals of about 100μV/m, the adjacent-channel signal need be only 0 to 10dB below the wanted signal. Thus there is little difficulty, and the geographical positioning of stations can easily be chosen so that adjacent-channel interference is substantially avoided.

F.M. Receivers

A number of types of British, American and continental commercial receivers have been used for listening to the Wrotham transmissions on f.m. Considerable differences in their performance have been observed, particularly for the all-important suppression of impulsive interference and receiver noise. If full advantage is to be taken of f.m., careful, but not necessarily costly, receiver design is necessary, particularly for second-class service areas. A good f.m. broadcast receiver should have:

(a) A limiter which can be set to operate efficiently at the minimum field strength for which the receiver is designed.

(b) An oscillator in which frequency drift is reduced to a reasonable minimum either by stabilizing crystal or some form of automatic frequency control. (In America, f.m. suffered some unpopularity due to receiver oscillator drift.)

(c) Some form of tuning indicator or even automatic tuning device is preferable. (The correct tune point is not easily identifiable as there is no point of maximum signal to which to tune.)

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For the conditions which will apply in the United Kingdom, the author suggests that the overall response of the h.f. and i.f. stages which precede the limiter should be not more than about 6dB down at ±85kc/s and about 20dB down at ±120kc/s. It is good practice to design all the interstage couplings, including those to the limiter and the discriminator, so that they are reasonably linear networks with respect to amplitude and phase each in their own right.

Such provisions perhaps tend to make f.m. receivers slightly more expensive than their a.m. counterparts, but nevertheless quantity production of crystals of adequate stability is practicable and manual tuning of a receiver does not present much difficulty. Incorrect tuning produces a degradation in quality little worse than that encountered on conventional m.f. receivers, but the performance of the f.m. receiver on noise suppression is also somewhat degraded. It can be expected that future developments will improve performance and reduce cost, but nevertheless it is to be hoped that f.m. receiver design will follow good technical practice and not omit a few circuit elements whose adjustment is not difficult in order to save a few shillings at the expense of the basic advantage of f.m., namely, improved quality with less interference.

The Wrotham High-power Experiment

The data obtained from the low-power trials indicated that full-scale tests using high-power transmissions provided by properly engineered service equipment would be justified, and at a site some 20 miles south-east of London the BBC constructed the Wrotham experimental v.h.f. station. Two transmitters were originally installed, a 25kW f.m. transmitter operating on a centre frequency of 91·4Mc/s, and a class-*B* modulated a.m. transmitter, using the same transmitting valve complement, giving an output power of about 18kW on a carrier frequency of 93·8Mc/s. Both transmissions were radiated simultaneously from a common aerial system having a gain of about 8dB and providing horizontally polarized radiation. The centre of the aerial system is some 1100ft above sea level and the maximum e.r.p. was approximately 150kW on f.m. and 110kW on a.m. Recently the a.m. transmitter has been converted to f.m., and a third 20kW f.m. transmitter, consisting of two 10kW units operated in parallel, is now being installed in readiness for the regular three-programme service which is scheduled to start from this station in May of this year. Fig. 3 shows the transmitter hall and the two original transmitters.

A second field strength survey on about 90Mc/s using both horizontal and vertical polarization was made from the Wrotham site using low power. This survey[3] provided the data for predicting the service area of the final station, gave additional information about the variations in field strength in built-up areas and over average ground contours, and assisted in formulating a method of estimating site-height and transmitting-aerial height-gain for v.h.f. transmitters[3]. Measurements in flat built-up districts showed that buildings of normal height for London reduce the average field, at a receiving-aerial height of 20ft, by about 10 to 12dB and that an average for one street will be the average for the district to an accuracy of about ±4dB. The variation along the street is less with horizontal than with vertical polarization. In hilly country the minimum field strength occurs on the near-side slope of a valley and not at the lowest point, and is frequently 30dB less than on flat ground on either side of the valley.

The minimum value is slightly lower for horizontal polarization than with vertical polarization. Measurements along a number of radials, each between 40 to 70 miles long, were used to prepare a predicted field strength contour map for the station and showed that with an operating power of about 125kW e.r.p., using f.m. or a.m.l. and either horizontal or vertical polarization, the station would give a satisfactory service to Greater London and most of South-east England. The predicted average field strength for almost the whole of London was 2mV/m or more, the greater part of Kent and Sussex also had a first-class service field, while for the coastal towns a field of between 250 to 100μV/m was predicted.

Experimental transmissions from Wrotham began early in 1950, and a comparison of the field strength of the station itself with that predicted showed that, while the results compared favourably, a slight change in site position can upset predictions if the ground section close to the transmitting aerial is different. The low-power tests used

Fig. 3. Transmitter hall at Wrotham

a site approximately half a mile east of the final aerial, and the observed differences in certain directions between the predicted and final measured values suggest that site tests on v.h.f. should be made on the exact site of the final transmitter if discrepancies of up to 10dB are to be avoided. The field strength map for the station operating at 125kW e.r.p. is reproduced in Fig. 4. It shows the average field strength and does not include the 10 to 12dB drops encountered in built-up areas.

Regular programme transmissions on both f.m. and a.m. began in July 1950 and a listening survey[8] was started. Listeners living within the projected service area were provided with receivers capable of receiving a.m. or f.m. and considered typical of a good commercial receiver of either type. Listening tests with a mobile laboratory were made in areas covered by the 100μV/m contour. For this survey two types of service area were defined as follows, and it was assumed that the listener's receiving aerial is within 30 to 60ft from a busy road on which traffic may be continuous:

(a) *First-class service area:* Impulsive interference for 50 per cent of motor-cars is imperceptible; of the remainder, occasional cars may produce interferences classed as " slightly disturbing."

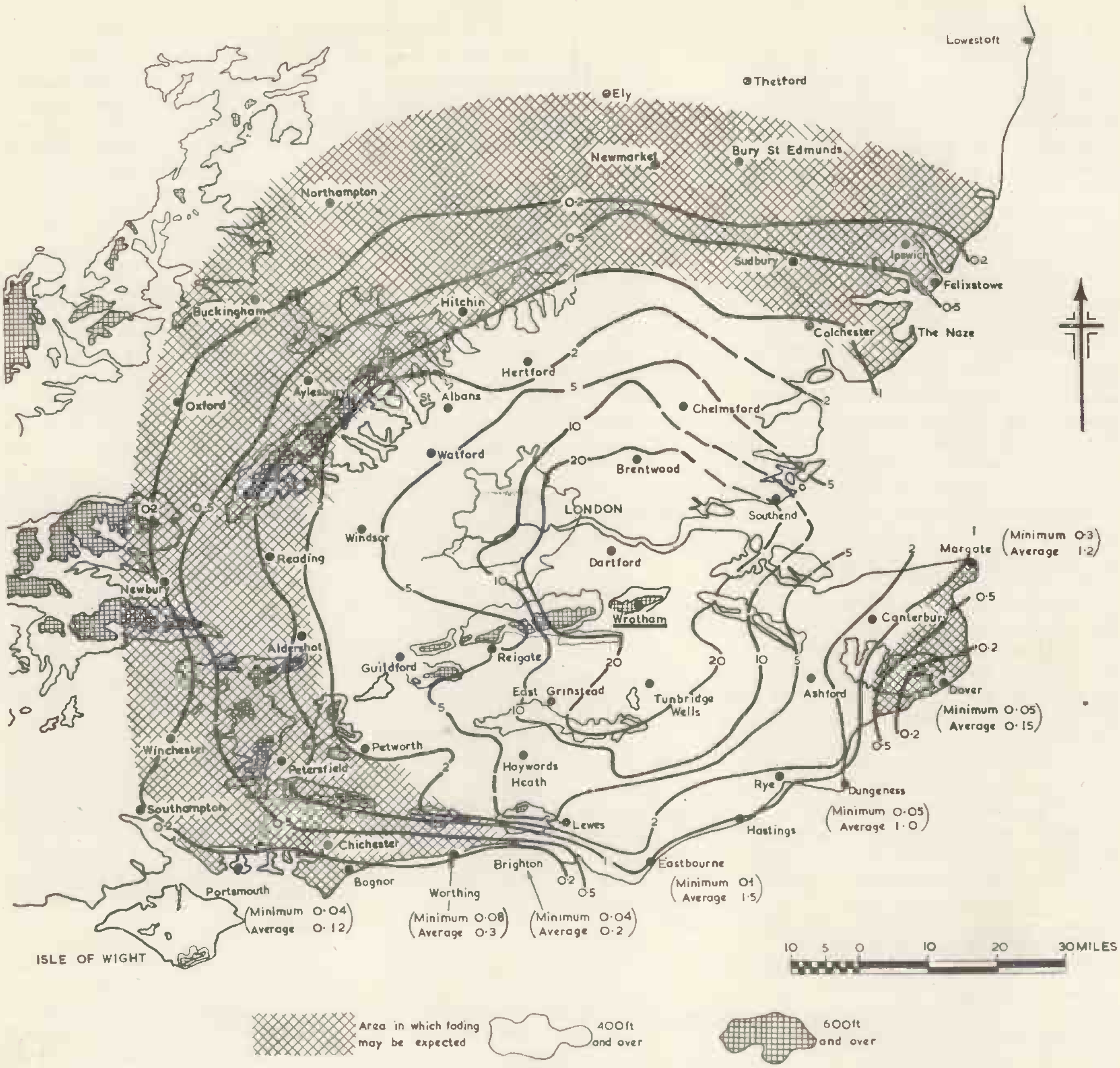

Fig. 4. Measured field strength contours of Wrotham v.h.f. transmitting station

Frequency 93·8Mc/s.
Effective radiated power 125kW.
Height of centre of aerial above sea level 1 100ft. approx.
Mast height 500ft.
Receiving aerial height 30 ft.
All contour numerals signify field strength in millivolts per metre.

(b) *Second-class service area:* The average level of interference from at least 50 per cent of motor-cars is never graded as worse than "perceptible." Occasional cars give rise to interference considered as "disturbing."

The results of the listening tests may briefly be summarized as follows:

1. Within 30 miles of the transmitter there was no general preference for either system of modulation; reception was excellent with a.m., a.m.l. and f.m. Any preference was for f.m. because of the absence of receiver hiss. Listeners were impressed by the simple aerial systems which gave satisfactory reception with f.m. and a large proportion of listeners could use an indoor aerial for f.m., which was not practicable on a.m.

2. Beyond 30 miles a majority of listeners expressed a distinct preference for f.m. Many listeners did not notice any marked difference between a.m.l. and a.m., but an improvement in favour of a.m.l. was admitted in areas subject to heavy impulsive interference. Occasional fading was noticed on a.m. and f.m. transmissions at distances in excess of 40 miles, and this was taken into account in assessing the service areas.

Table 1 shows the approximate average field strengths for the lower limit of the two types of service area defined and the corresponding approximate ranges from the transmitter. The coverage given by f.m. is seen to be greatly superior to that provided by a.m.l. or a.m. A general conclusion is that the use of f.m. can ensure a first-class urban service within the area bounded by the 1mV/m

TABLE 1.

SYSTEM	FIRST-CLASS SERVICE AREA		SECOND-CLASS SERVICE AREA	
	FIELD-STRENGTH (MV/M)	APPROX. RANGE (MILES)	FIELD-STRENGTH (MV/M)	APPROX. RANGE (MILES)
f.m.	1	45	0·25	60
a.m.l.	3	35	1	50
a.m.	10	25	3	35

contour and second-class service up to the limit of the 250μV/m contour. In areas remote from busy roads a first-class rural service could be given with 150μV/m, and a second-class service with 75μV/m.

It has been concluded that if a.m. were chosen the number of broadcasting stations required in the United Kingdom to give an equivalent service would be about three times as great as for f.m.[8].

The extra capital expenditure involved on sites, buildings and equipment if a.m. were used would be very heavy, particularly when it is remembered that the scheme envisaged by the BBC involves the distribution of three programmes (Home, Light and Third), so that each station needs three pairs of transmitters. The running costs due to staff, maintenance, electric power and the rental of high-grade programme lines would also be seriously increased. The result would be that in practice a compromise scheme giving less effective coverage would have to be accepted. From the point of view of a broadcasting authority, therefore, the technical and economic advantages of f.m. would appear to be quite clear.

V.H.F. Transmitters

For v.h.f. broadcasting the mode of wave transmission is virtually line-of-sight propagation from a radiator in free space, whereas for m.f. broadcasting ground-wave propagation from a radiator "on the ground" is used. The factors which determine the required e.r.p. of a broadcast station, to serve a given area, thus differ considerably for the two cases. Propagation at v.h.f. has been studied theoretically and experimentally, and the knowledge acquired is gradually being related to practical engineering. Even so, preliminary forecasts based on such data and on a survey of population densities and ground contours should finally be supplemented by actual field strength measurements around experimental transmitters set up on appropriate sites within the areas it is desired to serve. Only in this way can a sufficiently accurate estimate be made of the required e.r.p. for a particular station. Assuming f.m., the estimated e.r.p.'s of the 20 or so stations required by the v.h.f. service planned by the BBC range from about 10 to 120kW.

Transmitting aerials for m.f. broadcasting are physically restricted to single quarter-wave or half-wave radiators. At v.h.f., however, large aperture aerials consisting of many stacks of half-wave radiators are possible. Such aerials confine the greater part of their radiation to directions near the horizontal, and practical aerials for Band 2 may have overall net power gains of about 6 to 8 times that of an equivalent single half-wave dipole. The Wrotham Band 2 aerial, for example, has a gain of about 6. The production of a given e.r.p. (the product of the transmitter power and aerial gain) thus becomes an engineering and economic compromise between transmitter power and aerial construction. This can be a difficult choice to make, but the tendency, particularly in America, is to use the maximum convenient aerial system gain. Taking this as about 6 for Band 2 aerials, the powers of the transmitters themselves will range between about 2 to 20kW, appropriate intermediate values being 5 and 10kW. The transmitters for a v.h.f. service are thus in a lower power range, and hence are smaller and cheaper, than their m.f. equivalents.

Further practical simplification of the transmitters occurs if the system of modulation is f.m. The frequency modulated signal is of constant amplitude and is generated at a low power level. Thus, power amplification of the signal up to the final required value can be carried out in the transmitter with straightforward high-efficiency class-*C* r.f. amplifiers in cascade[9], and the relatively large audio frequency modulation amplifier required with conventional class-*B* modulated a.m. transmitters is disposed of. Each power amplifier can be arranged to work at its optimum continuous rating, while the final power amplifier carries only a quarter of the peak power of its high-level modulated a.m. equivalent. The amplitude-linearity characteristic of the amplifier stages is of no serious consequence, but the circuits must have an appropriate bandwidth (about 250kc/s) and a linear phase-frequency characteristic to minimize distortion.

The Wrotham f.m. transmitter was constructed at a time when the range of v.h.f. transmitting valves was very limited. Tetrodes give a high stage gain and are very suitable for the duty, but they were available in ratings suitable for the low-power amplifier stages only. Triodes of relatively low slope, in coaxial earthed-grid circuits, had to be used for the high-power amplifiers, and this resulted in a rather large number of stages because of the low stage gain (about 3·1) of this type of circuit when used with the triodes concerned. For future f.m. transmitters this could be avoided, because the present range of v.h.f. tetrodes and high-slope triodes has now been extended to include reliable valves having anode dissipations of 15kW or more for Band 2. An American v.h.f. tetrode of 10kW rating, now also being made in this country, is capable of giving a stage power gain of about 20 for frequencies up to 220Mc/s, while a new 10kW high-slope triode of continental origin is giving a power gain of about 10 at this frequency. Recent American developments in "inside-out" valves are enabling output powers of about 10kW at frequencies up to 900Mc/s to be obtained from a single valve with reasonable gain.

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(*To be continued*)

The Automatic Recording of Animal Behaviour in the Field

By R. J. Canaway*,
W. F. Raymond*
and J. C. Tayler*

This article describes a simple apparatus for the continuous non-subjective recording of animal movement under field conditions. The necessary switches are connected by cables to the self-contained recording unit which is carried on harness worn by the animal. Movement is unrestricted, and the animal can wear the equipment over long periods without discomfort.

WITH increasing world emphasis on the improvement of output from pastureland, a knowledge of the behaviour of grazing animals in relation to their productivity becomes of importance. In America, Johnstone-Wallace carried out early behaviour work with grazing beef cattle[1], and similar studies have been made in New Zealand[2], the United Kingdom, etc. Hughes and Reid[3] have recorded grazing and ruminating times of cattle and sheep at intervals through the season, and Tayler[4] has studied the effect of different systems of management on the behaviour and live-weight of bullocks. Based on studies of this type, Castle *et al.*[5] have suggested improvements in the grazing management of dairy cattle.

All these workers have recorded behaviour in the field by visual observations over periods of only a few days. The records taken, however, are seldom continuous, are subjective, tend to vary between different observers, and are difficult to take on numbers of animals at a time, on unrestricted grazing, or at night (to aid night observation, Castle *et al.*[5] have used infra-red equipment). While recording of behaviour by mechanical means has been carried out on animals in stalls, where careful adjustment of the equipment is possible, no equipment suitable for recording animal behaviour under the more rigorous conditions found in the field has been described.

In discussion with Dr. J. E. Lovelock of the National Institute for Medical Research in 1947, it was agreed that a method of automatically recording behaviour in the field should be possible. Preliminary tests were made by transmission of radio signals from a bullock, but it was evident that restriction of frequency allocations would limit the use of such equipment, and it was decided to work on portable recording equipment carried by the animals. The radio link, however, proved valuable in testing switches.

The equipment described here has been designed for use only on cattle: it will be obvious that many improvements are possible, but an attempt has been made to establish some of the basic requirements of such equipment.

Experimental

The recording box and switches are mounted on a leather harness worn by the animal. This has been developed from that described by Balch *et al.*[6], and can be worn for a period of several weeks without discomfort.

The most important agricultural information required includes the time spent by an animal in walking, standing, lying down, grazing and chewing the cud, together with the total number of leg and jaw movements. It was decided that this could be obtained by the use of switches operated by the following movements:

(a) *Walking:* a circuit actuated each time the animal takes a complete pace—this to operate on one leg only.

(b) *Lying-standing:* a switch operated by pressure of the body when the animal lies down.

(c) *Jaw switch:* an electric impulse to be set up by each jaw movement of the animal.

(d) *Head up-head down switch:* this switch to operate different circuits when the animal's head is up or down. A jaw movement (c) when the head is down to be recorded as a grazing movement, and with the head up as a cudding movement.

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These four switches to be connected to a recording box carried by the animal, from which a record of the previous 24 hours' activity can be taken each day. The photograph on page 102 shows a bullock carrying the equipment at present in use.

Switches

(a) Walking Switch

Various types of switches were tested, operated by pendulum (cf. pedometers), levers, mercury switches with internal oil-damping, etc. These were all liable to produce double-counting due to the uneven pace of a bullock when walking, and were frequently actuated by slight leg movements when the animal was not walking. These difficulties

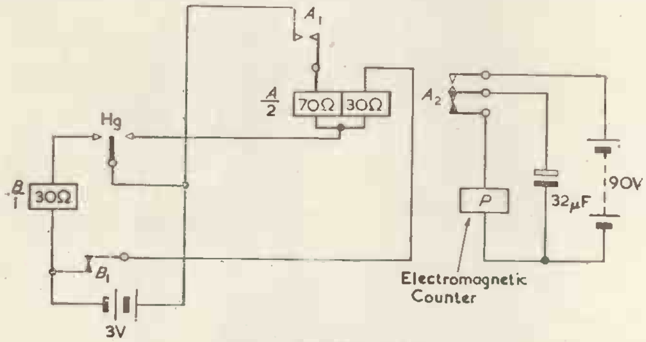


Fig. 1. Walking switch

Fig. 2. Construction of the walking switch

were overcome by the use of a single-pole changeover mercury switch, connected* to two relays (Fig. 1). Tilting of the switch by a leg movement operates relay A, and capacitor C is charged via contacts A_2. Relay A is now held in the on position by its hold winding. On completion of the leg movement, relay B is energized, so opening contacts B_1 and cutting off current supply to the hold winding on relay A. Relay A de-energizes, and capacitor C discharges through the counter. Thus a count is only made when a complete leg cycle occurs. The switch is mounted inside the lid of a box strapped just below one knee of the bullock (see photograph). The lid is mounted so that the mercury switch can be adjusted and locked in the position giving the most precise response to leg movements. In tests, this switch has given "walking" counts agreeing within 1 to 2 per cent with those obtained by visual recording. The construction of the switch is shown in Fig. 2.

* *An elastic telephone flex is used so as to avoid any slackness in this wiring.*

(b) Lying-Standing Switch

A switch operated by pressure of the animal's body appears most suitable. Various shapes of rubber pressure bags fitted in a belly strap have been tested: these were connected to a pneumatic switch operating when the pressure in a bag was increased by the animal lying on it. To allow for various lying positions, either a long bag (similar to a bicycle inner tube) or several smaller bags were necessary, and the system was found to be very sensitive to slight air-leaks. A more positive type of switch has been developed: in this the animal when lying down presses a spring-loaded plate mounted in a rubber cushion in the belly-band. This operates through a cam on to a micro-switch (Fig. 3). Two of these switch mechanisms are spaced along the band, and are wired so that pressure applied to one or both of them produces the same circuit changeover (Fig. 4). When the animal lies down the

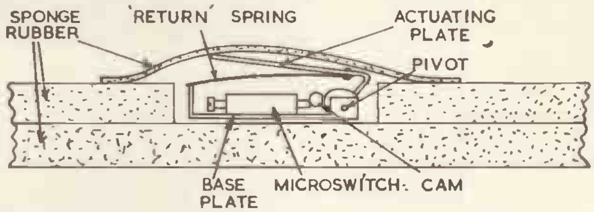


Fig. 3 (above). Standing-Lying down switch

The actuating plate forms a "lid" over the switch and protects it from damage due to the weight of the animal.

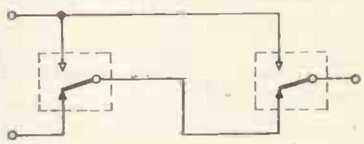


Fig. 4 (right). Standing-Lying down switching arrangement

Switches shown in the unoperated position.

switch is on: at the same time the walking (a) and head up-head down (d) switches are disconnected. These are not required when the animal is lying down, and this allows battery economy.

(c) Jaw Movements

This proved to be the most difficult movement to record precisely, because the forms of movement in grazing and cudding differ. Johnstone-Wallace[1] has used a pneumatic switch for recording jaw movements under indoor conditions. Various mechanical switches have been tested, but their differentials are most sensitive to slight changes in the position of the head-band in which they are mounted, which is far more difficult to keep adjusted on animals in the field than in the stall. Preliminary tests have shown a marked sound waveform related to jaw movements, which can be detected by a microphone held against the hard bone of the head. It may be possible to amplify this sound sufficiently to operate recording equipment, but problems of positive mounting of a microphone and suitable amplifying circuits have not yet been solved.

(d) Head up-Head down

This is a simple on-off switch operated by a cam mechanism from a loop arm resting on the neck of the animal: the arm is designed to operate even when the animal's head is turned sideways. The cam can be adjusted relative to the switch so that the latter is on

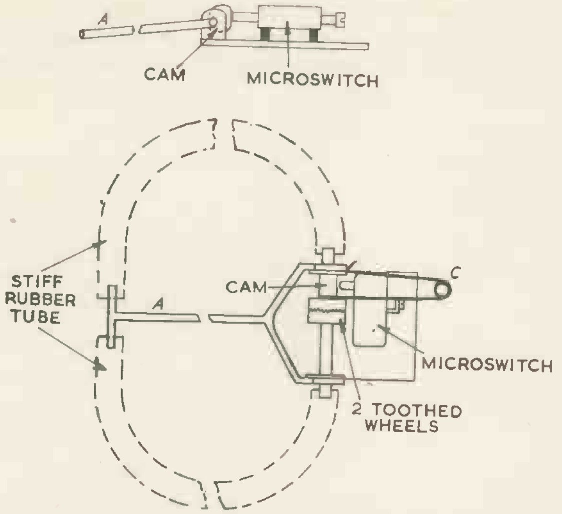


Fig. 5. Head up-Head down switch

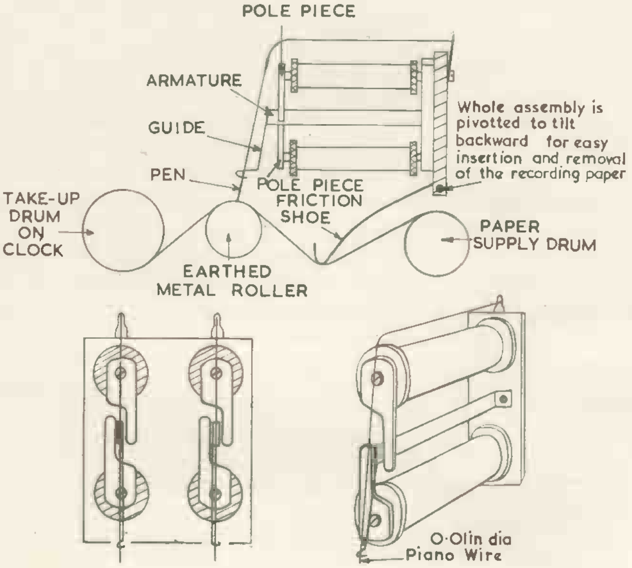
Spring clip C is accessible from the outside of the switch assembly and by pressing its arms together between finger and thumb, the two toothed wheels disengage. The arm A may then be quickly and easily reset relative to the operating point on the cam. Releasing the spring clip allows the toothed wheels to re-mesh and so lock the arm and cam in the correct position.

only when the animal's head is low enough for it to graze (Fig. 5).

Method of Recording

Two types of record are required, namely, the distribution of various activities within the day, together with their intensity and duration, and also the total number of certain movements. For the latter a simple electromagnetically operated counter seems satisfactory, but various methods of recording "activity" are possible. It was decided this could best be made on a strip of moving paper taken up on a constant-speed drum: recording on this paper could either be photographic, by a stylo on kymograph or waxed paper, by ink on paper, or electrically on Teledeltos paper. After various tests the latter method was adopted. A 48in length of paper per 24 hours gives a reasonable discrimination of activities, and is short enough to allow daily "writing up" of the observations.

Fig. 6. Pen mechanism general arrangements



The paper recorder contains a 3in take-up drum, operated by a clock mechanism, which draws a ½in wide strip of Teledeltos paper across an earthed metal roller, with two pens resting on the upper surface of the paper. The paper is taken from a spool which is replenished at intervals.

After initial tests it was found the clearest and most reliable record would be obtained by having a continuous trace (i.e. h.t. applied continuously to the pens), movements of the animal being indicated by lateral movements of the pens across the paper. This has been done by con-

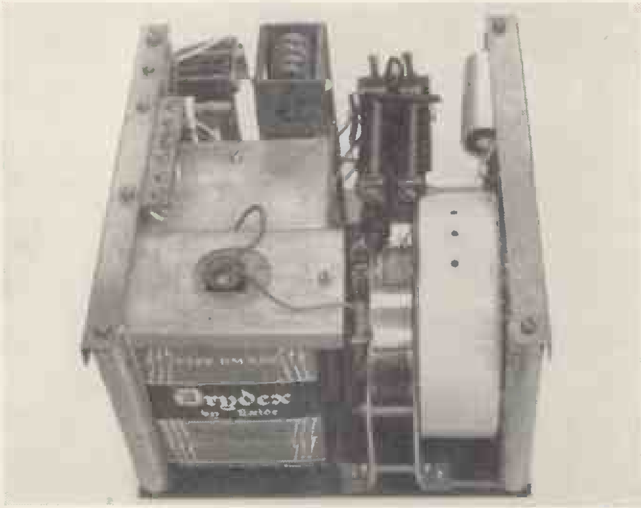


Fig. 7. Layout of the pen and relay unit

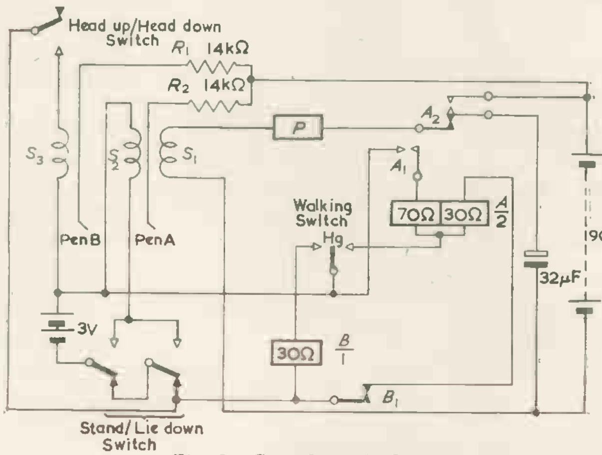


Fig. 8. General circuit diagram

trolling the positions of the pens by solenoids, which are operated, in some cases via relays, from the various body switches (Fig. 6). (The use of l.t. relays is desirable, to avoid h.t. outside the recording box.) The pens take an average current of 3 to 5mA at 90V, and the solenoids operate on 50mA at 3V. Current passes continuously between each pen and the earthed metal roller, through the Teledeltos paper, producing a blackening of the paper. As the paper moves, two black lines are produced on it: lateral movement of the pens, actuated by various body movements, make "blips" on these lines.

The two pens, paper mechanism, relays, etc., are enclosed in a rigid metal box, approximately 6in × 6in × 4in. The layout (Fig. 7) is designed to allow easy removal of the paper record and replacing of batteries. The present boxes weigh 5lb, but a reduction in weight is planned in further development. Connectors join the flexes from the various body switches to the recorder box, and these can readily be broken so as to allow the box to be removed. The box is mounted on a saddle on the bullock's back.

The full circuit for the body switches and recorder is shown in Fig. 8.

When the animal is standing, pen *A* produces a horizontal trace on the paper. Each complete leg movement momentarily energizes solenoid S_1, causing the pen to move vertically across the paper. When the animal is lying down, a second solenoid S_2 pulls the pen in the opposite direction from S_1. Thus the trace from pen *A* is interpreted as in Fig. 9.

Because of the slow movement of the paper, the total

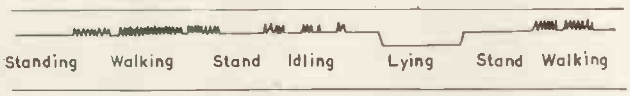


Fig. 9. Record produced by pen A

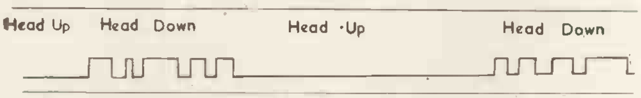


Fig. 10. Record produced by pen B

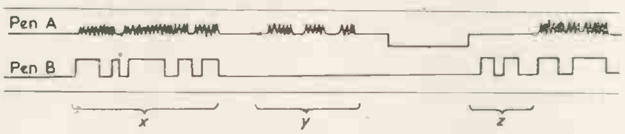


Fig. 11. Record produced by pens A and B together

number of leg movements cannot be counted from the trace, and is found from the counter *P*. This gives the total leg movements per 24 hours: the distribution of this activity during the day is found from the duration and degree of darkening on the trace (e.g. distinction between walking and idling in Fig. 9).

Pen *B* is controlled by the head up-head down switch, through solenoid S_3, and produces a trace (Fig. 10).

When the development of switch (c) is completed, jaw movements will be superimposed on this trace. However, much information on the activity of grazing is obtained by study of the two traces from pens *A* and *B*, which are produced side by side on the paper (Fig. 11).

Thus, walking with head down is taken as grazing (x), walking with head up as idling (y), while standing still, even though the head is down (z), is unlikely to be grazing.

The rate of movement of the paper strip increases slightly during the 24 hours, due to its being taken up on a drum revolving at constant speed. Interpretation of the data is facilitated by holding the strip from each day alongside a 48in rule calibrated directly in hours, so as to allow for non-linear movement of the paper.

While the present equipment has been used successfully in several experiments, further development and improvement are desirable. Thus a single 90V layer-type battery will only supply power to one recorder for 48 hours (the 2 × 1½V l.t. batteries will operate for a week), and methods of battery economy are being studied, together with the possibility of reducing the size and weight of the recording box. The main development outstanding is the design of suitable jaw recording equipment, but it is evident that useful data on grazing can be obtained without this.

Method of Use

The bullocks on which the equipment is to be used must be accustomed to wearing the harness, first with a dummy recording box on the back, and then to the switches on the body. This may take a week or longer, but the bullocks we have used have all lost their initial restiveness in the harness, though they never seem to lose a liking for chewing leather, flexes, etc. However, with experience, damage of this sort can be minimized.

Rubbing by the animal against gates and fences can be more serious, and at least in the early stages of this work the use of electric fencing, which offers no projections for rubbing, is preferable. After adapting the animal to the harness, etc., two or three days are necessary for adjusting the exact positions of the leg and neck switches: at this stage the traces and counts are checked against visual observations. When these are satisfactory, the equipment can be used to obtain behaviour data of the animals under various grazing conditions: the used length of paper strip is removed every 24 hours, and the counter figure read.

Preliminary Use of the Recording Equipment

A pair of identical twin Hereford-cross bullocks of quiet disposition was used for initial testing of the equipment described. It was found that, provided the harness was correctly adjusted, behaviour of a harnessed twin was not adversely affected in comparison with its twin in the same paddock without harness (cf. Hancock[2]). In later tests, both twins were harnessed and carried recording boxes. The recording mechanism withstood the shaking due to walking and running of the bullocks and gave satisfactory operation over a number of consecutive days. Adjustments were required chiefly in the external attachment of the various switches, due to the difficulty of adjusting straps to be firm but not uncomfortably tight. Excellent agreement was found between observations of up to an hour in which movements were recorded in detail by a stop-watch and the record subsequently " read " from the paper strip for that period.

Acknowledgments

The authors' thanks are due to the Avon Rubber Company, Melksham, Wilts; Burgess Products Company Ltd., Gateshead; Bradbury Brothers, Reading; Drayton Thermostat Company, Middlesex; and Recorder Charts Ltd., London, for advice and help with equipment used. Also to Mr. Pinniger, National Institute for Research in Dairying, Shinfield, for making the leg-switch cases, and to Dr. William Davies and other colleagues for their encouragement of this work.

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The Operation and Loading Characteristics of Valve Oscillators for Dielectric Heating

(Part 1)

By V. L. Atkins*

This article gives a mathematical analysis of valve oscillators as used in dielectric heating. The analysis is based on class-C amplifier operation. The conditions for maximum efficiency are shown and the effects of variable loading on the oscillator are dealt with. Later sections of this article will deal with power oscillator circuits, load matching and power transfer and work circuits and the control of loading.

DURING the last few years a considerable amount of interest has been aroused in industry by the use of high frequency power to heat materials of low electrical conductivity. This relatively new method of heating which differs from that of the normal conventional heating processes of convection, conduction and radiation, in that the heat is generated within the mass of material itself and does not depend upon heat transfer, has become of increasing importance to the development of new and improved methods of production. It has enabled production costs to be lowered, quality to be improved and has been the means of solving certain difficult heating operations. The material is heated when it is subjected to a high frequency alternating electric stress, the heat being generated by intermolecular orientations and ionic oscillations set up within the structure of the material. Provided the field distribution is uniform and the material homogeneous, the heat generated throughout the mass will be uniform, being directly proportional to the frequency and to the square of the applied voltage. The rate of heating also depends upon the electrical and physical properties of the material, the heating ability of a dielectric being measured in terms of its loss factor $\epsilon \tan\alpha$.

The only suitable device which can supply the power at the frequencies required for dielectric heating is the thermionic valve. In most types of industrial high frequency equipment designed for dielectric heating applications, the power is generated by operating either a single or a number of triode power valves as a self-excited oscillator unit. The material to be heated is arranged to form the dielectric of a capacitor which is either the main tank capacitor of the oscillator or is connected to the tank circuit by various coupling arrangements.

One of the main problems associated with this type of equipment is that its operation tends to be rather critical and unless the correct operating conditions are obtained, the performance of the equipment can be seriously affected. This calls for a considerable amount of thought and ingenuity in design, if satisfactory performance of the equipment is to be obtained under conditions which are liable to be met in practice. In most applications the load imposed on a generator is likely to vary during a heating cycle, and unless the generator is equipped with means of controlling the load, the power output can be seriously reduced. The problem is increased still further in the case of equipment required to cover a wide range of applications of different loading characteristics. Discretion must be used in deciding to what extent the controlling circuit may be taken, for if it is allowed to become too complex, then the additional costs involved and skilled operation required, may not justify the use of the equipment. In certain single application work, means of automatically matching the load to the generator have been developed and so permit ease of control over the equipment. On the other hand it may be found that the amount of variation in power output which occurs during a heating cycle for a particular material is not so serious as to warrant adjustments of loading. The problem is greatly simplified where the heating process can be carried out on a continuous basis, as the load will then present some average but constant value which can be adjusted to the optimum value. In order to be able to select the most suitable operating conditions for a particular application and use the equipment to its best advantage, it is necessary to analyse the performance of the oscillator under varying load conditions. This first requires an analysis of valve operation to be made.

Furthermore due to the severe usage a valve is liable to meet with in industrial equipment, it is normal practice to derate the operating conditions as specified for when the valve is operated under ideal conditions such as obtained in communication and broadcasting installations.

Analysis of Valve Operation

The behaviour of a power oscillator depends upon numerous interdependent variables, all of which govern the power output and operating efficiency; the carrying out of a detailed investigation of a valve's performance under varying operating conditions presents a formidable task. However, certain approximations can be made which allow sufficiently accurate results to be obtained for most practical design work. The whole design problem revolves around obtaining the correct operating conditions to ensure the correct loading of the valve to deliver maximum power output, sufficient grid excitation to draw the full rated peak emission of the cathode and the correct level of d.c. bias to give a high efficiency of operation. The operation of an oscillator does not differ greatly in principle from that of a class-*C* amplifier, the main difference being that, in the case of an oscillator, the power absorbed by the grid circuit is taken from the output circuit. Thus in the following analysis which is based on class-*C* amplifier operation, the grid driving power must be deducted from

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the total power output. To simplfy the computations involved, it is assumed that the tank circuit always presents a resistive load to the valve and the frequency of operation does not vary greatly with load.

Voltage Current and Power Relations

The anode and grid voltage and current relations of a valve operating as a class-*C* amplifier are shown in Fig. 1. The instantaneous values of anode and grid currents depend upon the combined action of the voltages appearing at the anode and the grid at a particular instant and may be evaluated from the characteristic curves of the valve.

The voltage which appears at the anode of the valve consists of the d.c. supply voltage V_o minus the fundamental alternating voltage V_a which is developed across the tank circuit. Under normal operating conditions the peak amplitude V_a of the fundamental approaches to approximately 90 per cent of the direct voltage V_o, and thus the minimum instantaneous anode voltage $v_{a(min)} = (V_o - V_a)$ will approximate to 10 per cent of V_o.

The voltage appearing at the grid consists of an alternating voltage V_g (peak) superimposed on a negative d.c. bias voltage V_B which is greater than the cut-off value of the valve V_c. The phase relations of anode and grid voltages are such that the minimum instantaneous anode voltage $v_{a(min)}$ occurs at the same time as the maximum instantaneous grid voltage $v_{g(max)}$ and corresponds to the point when the anode and grid currents reach maximum values. The sum of the anode and grid currents constitutes the total space current which is emitted by the cathode. The peak value of the space current which can be permitted to flow is governed by the emission characteristic of the cathode and determines its useful life.

When operating correctly the value of the maximum grid voltage $v_{g(max)}$ is slightly less than the value of the anode voltage $v_{a(min)}$. If the grid is driven positive with respect to the anode (i.e. $v_{g(max)} > v_{a(min)}$) then the valve is likely to be damaged due to the excessive grid current which will be drawn.

The instantaneous anode and grid voltages can be expressed as follows:

Instantaneous anode voltage:

$$v_a = V_o - (V_o - v_{a(mim)}) \cos\beta \quad \ldots\ldots \quad (1)$$

Instantaneous grid voltage:

$$v_g = (V_B + v_{g(max)}) \cos\beta - V_B \quad \ldots\ldots \quad (2)$$

where β is the number of electrical degrees measured from the crest of the cycle.

List of Symbols Used

$f_1(\theta_a)$	$= \dfrac{\theta_a - \sin\theta_a \cos\theta_a}{\sin\theta_a - \theta_a \cos\theta_a}$
$f_2(\theta_a)$	$= \dfrac{\pi(1 - \cos\theta_a)}{\sin\theta_a - \theta_a \cos{}_a\theta}$
$f_3(\theta_a)$	$= \dfrac{\pi}{\sin\theta_a - \sin\theta_a \cos\theta_a}$
$f_4(\theta_a)$	$= f_3(\theta_a)(1 - \cos\theta_a)$
g_m	Mutual conductance.
I_a	Peak amplitude of the fundamental a.c. component of the anode current (output).
$I_{a(act)}$	Actual value of a.c. component.
$I_{a(opt)}$	Optimum value of a.c. component.
I_g	Peak amplitude of the fundamental a.c. component of the grid current.
$I_{g(dc)}$	D.C. grid current.
I_k	Cathode current.
I_o	D.C. anode current (input).
$I_{o(act)}$	Actual value of d.c. current.
$I_{o(opt)}$	Optimum value of d.c. current.
i_a	Instantaneous anode current.
$i_{a(pk)}$	Peak amplitude of anode current.
$i_{g(pk)}$	Peak amplitude of grid current.
L	Tank circuit inductance.
$L_{(opt)}$	Optimum value of inductance.
P_a	Power dissipated at anode.
P_g	Power dissipated at the grid electrode.
$P_{g(tot)}$	Total grid driving power.
P_{in}	Power input to valve.
P_{out}	Power output of valve.
P_{Rg}	Power dissipated in grid resistance.
Q	Q of tank circuit.
$Q_{(act)}$	Actual Q value.
$Q_{(opt)}$	Optimum Q value.
R_g	Grid resistance.
R_L	Effective parallel resistance of anode load.
$R_{L(act)}$	Actual load resistance.
$R_{L(opt)}$	Optimum load resistance.
R_o	Dynamic internal impedance of oscillator $(= f_3(\theta_a)\, r_a)$.
R_s	Effective series resistance of tank circuit.
r_a	Internal impedance of valve.
V_a	Peak amplitude of the fundamental a.c. voltage developed across the anode load.
V_B	Bias voltage.
V_c	"Cut-off" voltage.
V_g	Peak amplitude of grid exciting voltage.
V_g'	Peak amplitude of grid exciting voltage under dynamic conditions.
V_o	D.C. anode voltage.
V_q	Voltage by which bias exceeds "cut-off" value.
v_a	Instantaneous anode voltage.
$v_{a(min)}$	Minimum instantaneous anode voltage.
v_g	Instantaneous grid voltage.
$v_{g(max)}$	Maximum instantaneous grid voltage.
α	Ratio of $\mu V_B / V_c$.
Tan α	Loss factor.
β	Number of electrical degrees measured from the crest of the cycle.
γ	Ratio of $\mu V_{g(max)} / V_o$.
ϵ	Dielectric constant.
η	Valve operating efficiency.
$\eta_{(a)}$	Asymptotic efficiency.
$\eta_{(sat)}$	Efficiency at saturation.
θ_a	$\frac{1}{2}$ the angle of anode current flow (operating angle).
$2\theta_{a(act)}$	Actual operating angle.
$2\theta_{a(opt)}$	Optimum operating angle.
θ_g	$\frac{1}{2}$ the angle of grid current flow.
μ	Amplification factor of value.
ρ	$\dfrac{1 + \mu}{r_a}$
σ	Slope of limiting edge of a valves characteristics.
τ	$= \dfrac{v_{a(min)}}{V_o}$
ϕ	Ratio of R_L / r_a.
Ψ	Universal function.
ψ	Function of θ_a.

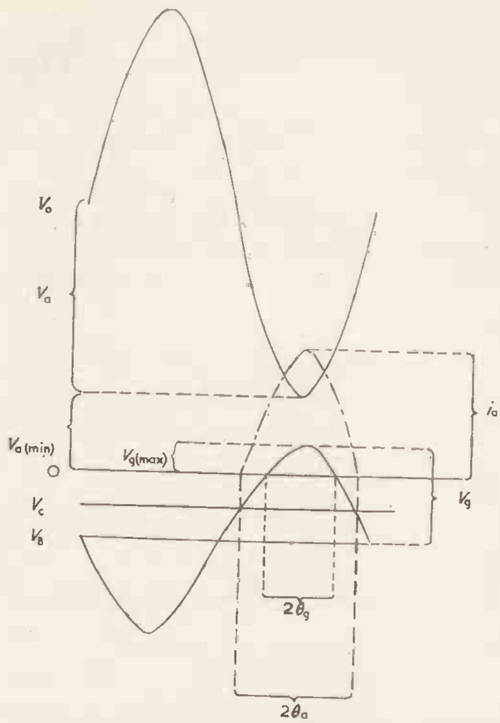


Fig. 1. The anode and grid voltage and current relations of a class-C amplifier or oscillator

During the positive swing of the grid voltage both anode and grid currents will flow. These currents are in the form of pulses, the duration of which are determined by the number of degrees by which the positive cycle of the grid voltage exceeds the cut-off point in the case of the anode current pulse, and the cathode (zero) potential in the case of the grid current pulse.

These angles of flow can be expressed as follows:

Angle of grid current flow equals $2\theta_g$ where

$$\theta_g = \cos^{-1} V_B/V_g \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

Angle of anode current flow equals $2\theta_a$ where

$$\theta_a = \cos^{-1} \frac{V_B - V_c}{V_g} \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

Equating equations (1) and (2) at the point of cut-off (V_c) where v_g is equal to v_a/μ, μ being the amplification factor of the valve, a further expression for the value of θ_a which eliminates the cut-off value V_c is obtained:

$$\theta_a = \cos^{-1} \frac{1}{1 + \dfrac{\mu\, v_{g(max)} + v_{a(min)}}{\mu V_B - V_o}} \quad \ldots\ldots\ldots \quad (5)$$

Simplifying this relation by taking the limiting case where $v_{a(min)}$ equals $v_{g(max)}$ and calling the ratio of this value to V_o, τ; the following useful expression is obtained:

$$V_B/V_o = 1/\mu \left[1 + \tau(1 + \mu) \frac{\cos\theta_a}{1 - \cos\theta_a} \right] \quad \ldots\ldots \quad (6)$$

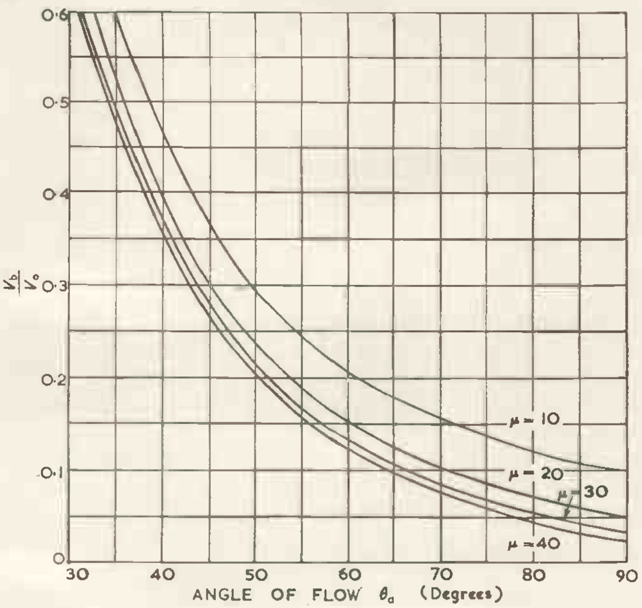
Curves showing values of V_B/V_o for varying operating angles and amplification factors for a value of τ equal to 0·1, are plotted in Fig. 2.

The power dissipated at the anode at any instant is equal to the product of the instantaneous values of anode current and voltage, while the useful power supplied to the tank circuit at any instant is equal to the product of the instantaneous values of anode current and the voltage developed across the tank circuit.

The maximum power which can be safely dissipated at the anode determines the power handling capacity of the valve. The power loss and output taken over a complete cycle is equal to the average of the respective instantaneous values. The sum of the power loss and output constitutes the power input to the valve, this being equal to the product of the anode d.c. supply voltage V_o and the direct current I_o which is the average value of anode current flowing over a complete cycle, as would be indicated by a direct current meter connected in the anode supply. The operating efficiency is the ratio of power output to input and depends largely upon the angle of flow $2\theta_a$. The high efficiency at which a class-C amplifier or oscillator operates is the result of the anode current flowing for something less than half a cycle, during which period most of the d.c. supply voltage is used across the tank circuit, whereas the anode voltage is passing through its minimum value. If the angle of current flow $2\theta_a$ is increased, the d.c. input will increase with the result that the power input, output and loss will increase. At the same time a greater proportion of the d.c. supply voltage acts upon the anode while the anode current flows and so reduces the efficiency. This means that a high efficiency of operation is only obtained at the expense of a low output power. A compromise therefore has to be made between output power and efficiency, and in practice efficiencies of the order of 60 to 75 per cent are obtained with durations of anode current flow of between 120° and 150°.

The pulse of anode current is distorted from being a section of a sine wave, owing to the valve operating over the non-linear portions of the characteristic curve which exists at saturation and cut-off. Furthermore the flow of grid current diverts some of the space current from the anode flow and thus tends to "flatten" the crest of the anode pulse. This distortion increases the harmonic content of the anode current waveform. However, as the impedance presented by the tank circuit to the fundamental frequency is considerably greater than the impedance offered to the harmonics, the predominating voltage developed across the circuit will be that of the fundamental.

Fig. 2. Ratio of $\frac{V_B}{V_o}$ against angle of flow θ_a for varying value of μ at the limiting conditions where $v_{a(min)} = v_{g(max)}$, having been taken to be equal to 0·1



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Thus the anode current will consist primarily of the d.c. component I_o plus an alternating current I_a of the fundamental frequency. The voltage developed across the tank circuit will be equal to the product of I_a and the effective load resistance R_L, i.e.:

$$V_a = I_a R_L \quad \ldots\ldots \quad (7)$$

As the values of I_a and V_a are the peak values then the power output (i.e. developed in the tank circuit) will be expressed as follows:

$$P_{(out)} = \frac{V_a I_a}{2} \quad \ldots\ldots \quad (8)$$

The power input:

$$P_{in} = V_o I_o \quad \ldots\ldots \quad (9)$$

and the valve efficiency:

$$\eta = \frac{V_a I_a}{V_o I_o} \ . \ 50 \text{ per cent} \quad \ldots\ldots \quad (10)$$

The resistive load R_L which the tank circuit presents to the valve is equal to $QL\omega$ and thus power output can be expressed as:

$$P_{(out)} = \frac{V_a^2}{2\,QL\omega} \quad \ldots\ldots \quad (11)$$

from which the following useful relation is obtained:

Effective voltage across tank circuit

$$V_{a(r.m.s.)} = \sqrt{(P_{(out)}\,QL\omega)} \quad \ldots\ldots \quad (12)$$

Analysis of Anode Current Components

The power input and efficiency of a valve operating under class-*C* conditions is governed largely by the angle of anode current flow $2\theta_a$, which determines the relation between the d.c. (I_o input current) and the fundamental a.c. (I_a output current) components of the anode current. This relation between the ratio of I_a/I_o and the operating angle $2\theta_a$ forms the basis upon which the operation of a valve working under different loading conditions can be analysed.

Under dynamic operating conditions the anode current flowing at any instant is a function of both the instantaneous values of anode and grid voltages, i.e.:

$$i_a = f\,.\,(v_g, v_a) \quad \ldots\ldots \quad (13)$$

Childs-Langmuir's "space charge formula" shows that the total current flowing from the cathode is proportional to the 3/2 power of the voltage (V) producing the electric field at the surface of the cathode, i.e.:

$$I_k = MV^{3/2} \quad \ldots\ldots \quad (14)$$

M being a constant depending upon the valve construction. In applying this relation to a triode the electric field existing at the surface of the cathode is proportional to the quantity $(V_g + V_a/\mu)$ which represents the effective anode voltage produced by the combined action of both grid and anode voltages. For most triodes the amplification factor μ is constant over their normal working range and the cathode current can be expressed as follows:

$$I_k = M(V_g + V_a/\mu)^{3/2} \quad \ldots\ldots \quad (15)$$

For the purpose of analysing the operation of a valve amplifier it is only the anode current relationship which requires to be studied, and in practice it is sufficient to assume that the relationship follows a linear law which simplifies the calculations involved. Actually, operating curves derived from the 3/2 power law do not differ greatly from those based on a linear law.

The anode current can be expressed as follows:

$$i_a = g_m(v_g + v_a/\mu) \quad \ldots\ldots \quad (16)$$

where g_m, the mutual conductance of the valve, represents the slope of the i_a/v_a characteristics of the valve (assumed to be linear). With class-*C* operation, where anode current only flows for part of a complete cycle the relation becomes:

$$\left.\begin{array}{ll} i_a = g_m(v_g + v_a/\mu), & v_g + v_a/\mu > 0 \\ i_a = 0, & v_g + v_a/\mu < 0 \end{array}\right\} \quad (17)$$

the change over from one equation to the other occurring when $v_g = -v_a/\mu$ i.e. when $\beta = +\theta_a$ and $-\theta_a$.

Substituting for the values of v_a and v_g as given by equations (1) and (2):

$$i_a = g_m\left[(V_B + v_{g(max)})\cos\beta - V_B + V_o/\mu - \frac{V_o - v_{a(min)}}{\mu}\cos\beta\right] \quad \ldots\ldots \quad (18)$$

Under static conditions the "cut-off" point is equal to V_o/μ and putting

$$V_c = V_o/\mu = V_B - V_q \quad \ldots\ldots \quad (19)$$

where V_q is equal to the voltage by which the bias exceeds the cut-off value, then the expression for anode current becomes:

$$i_a = g_m(V_g' \cos\beta - V_q) \quad \ldots\ldots \quad (20)$$

and

$$\cos\theta_a = V_q/V_g' = \frac{V_B - V_o/\mu}{V_g'} \quad \ldots\ldots \quad (21)$$

where

$$V_g' = V_B + v_{g(max)} - \frac{(V_o - v_{a(min)})}{\mu} \quad \ldots\ldots \quad (22)$$

and represents the peak amplitude of the effective grid voltage which would produce the same change in anode current under static (constant anode voltage) conditions. The peak amplitude of the anode current will be:

$$i_a \text{ peak} = g_m V_g'(1 - \cos\theta_a) \quad \ldots\ldots \quad (23)$$

The instantaneous anode current which consists of the d.c., the fundamental a.c. and various harmonic components can be expressed as a Fourier series, i.e.:

$$i_a = f(\beta) = I_o/2 + \sum_{n=1}^{\infty}(I_{an}\cos n\,\beta + I_{an}\sin n\,\beta) \quad \ldots \quad (24)$$

Expansion of this series for the interval $-\pi$ to $+\pi$ gives:

$$i_a = f(\beta) = 1/2\pi\int_{-\pi}^{\pi} f(\beta')d\beta' + 1/\pi\sum_{n=1}^{\infty}\int_{-\pi}^{\pi}(f(\beta')\cos n\,\beta'\,d\beta')\cos n\,\beta$$
$$+ 1/\pi\sum_{n=1}^{\infty}\int_{-\pi}^{\pi}(f(\beta')\sin n\,\beta'\,d\beta')\sin n\,\beta \quad \ldots\ldots \quad (25)$$

In this expression the variable of integration in the definite integrals is written as β' to show that the coefficients are not a function of the variable of integration.

Extracting the values of the d.c. and the in-phase (cosine term) fundamental ($n = 1$) a.c. coefficients:

$$I_a = 1/\pi\int_{-\pi}^{\pi} f(\beta')\cos\beta' d\beta' \quad \ldots\ldots \quad (26)$$

and:

$$I_o = 1/2\pi\int_{-\pi}^{\pi} f(\beta')\,d\beta' \quad \ldots\ldots \quad (27)$$

Now i_a is a function of β thus:

$$I_a = 1/\pi\int_{-\pi}^{\pi} i_a \cos\beta\,d\beta \quad \ldots\ldots \quad (28)$$

and:

$$I_o = 1/2\pi\int_{-\pi}^{\pi} i_a \cos\beta \quad \ldots\ldots \quad (29)$$

Applying the values given for i_a in equations (17) and (20):

$$I_a = 1/\pi \left[\int_{-\pi}^{-\theta_a} 0 \cos \beta \, d\beta + \int_{-\theta_a}^{+\theta_a} g_m(V_g' \cos \beta - V_q)\cos \beta \, d\beta + \int_{+\theta_a}^{\pi} 0 \cos \beta \, d\beta\right] \quad \ldots\ldots\ldots\ldots (30)$$

and:

$$I_o = 1/2\pi \left[\int_{-\pi}^{-\theta_a} 0 \, d\beta + \int_{-\theta_a}^{+\theta_a} g_m(V_g' \cos \beta - V_q) \, d\beta + \int_{+\theta_a}^{\pi} 0 \, d\beta\right] \quad \ldots\ldots\ldots\ldots\ldots\ldots (31)$$

Owing to the symmetry, the equations become:

$$I_a = 2/\pi \int_0^{\theta_a} g_m(V_g' \cos \beta - V_q) \cos \beta \, d\beta \quad \ldots\ldots (32)$$

and:

$$I_o = 1/\pi \int_0^{\theta_a} g_m(V_g' \cos \beta - V_q) \, d\beta \quad \ldots\ldots\ldots (33)$$

Integrating and substituting the value of V_q as given in equation (21):

$$I_a = \frac{V_g' \, g_m}{\pi}(\theta_a - \sin \theta_a \, \cos \theta_a) \quad \ldots\ldots\ldots (34)$$

$$I_o = \frac{V_g' \, g_m}{\pi}(\sin \theta_a - \theta_a \cos {}_a\theta) \quad \ldots\ldots\ldots (35)$$

The ratio of the direct current to the fundamental a.c. component is now readily obtained, viz:

$$I_a / I_o = \frac{\theta_a - \sin \theta_a \cos \theta_a}{\sin \theta_a - \theta_a \cos \theta_a} = f_1(\theta_a) \ldots\ldots\ldots (36)$$

Thus the ratio of I_a / I_o is a function of the duration of anode current flow. Values are plotted in Fig. 3 for operating angles between 100° and 160° which is the usual range met in practice.

A further relation which enables the ratio of the peak current drawn by the anode to the average d.c. value to be evaluated is derived from equations (23) and (35), i.e.:

$$i_{a(peak)} / I_o = \frac{\pi(1 - \cos \theta_a)}{\sin \theta_a - \theta_a \cos \theta_a} = f_2(\theta_a) \quad \ldots\ldots (37)$$

Values for this relation are also plotted in Fig. 3.

The value of $i_{a(peak)} / I_a$ is simply obtained from $f_2(\theta_a)/f_1(\theta_a) = f_4(\theta_a)$.

Grid Driving Power

The flow of grid current causes power to be dissipated in the grid circuit, which in the case of a self-excited oscillator is drawn from the output power, which thus reduces the amount of useful power available for work application. The total power absorbed in the grid circuit is equal to:

$$P_{g(tot)} = \frac{I_g \, V_g}{2} \quad \ldots\ldots\ldots\ldots\ldots (38)$$

where V_g is the amplitude of the grid exciting voltage and I_g is the peak amplitude of the fundamental alternating current component. Part of this power is dissipated in the grid resistor (R_g) and will equal:

$$P_{Rg} = I_{g(dc)}^2 \, R_g \quad \ldots\ldots\ldots\ldots\ldots\ldots (39)$$

where $I_{g(dc)}$ is the direct current component; i.e. the average value as would be indicated by a d.c. meter connected in series with the resistor; while the remaining amount of the power is lost through heating of the grid electrode by electron bombardment. Under normal operating conditions the grid current flows only for a very small fraction of the complete cycle and as a result the ratio of the alternating component to that of the direct component approximates to 2. This can be easily seen, for as θ_g approaches zero the definite integral of the fundamental component coefficient approaches to the value of the definite integral of the direct component coefficient, i.e.:

$$\int_0^{\theta_g \to 0} i_g \cos \beta \, (d\beta) \to \int_0^{\theta_g \to 0} i_g \, (d\beta) \quad \ldots\ldots\ldots\ldots (40)$$

from which follows:

$$I_g / I_{g(dc)} = \frac{2/\pi \int_0^{\theta_g \to 0} i_g \cos \beta \, (d\beta)}{1/\pi \int_0^{\theta_g \to 0} i_g \, (d\beta)} \to 2 \quad \ldots\ldots (41)$$

thus the total driving power can be expressed as:

$$P_{g(tot)} = I_{g(dc)} \, V_g \quad \ldots\ldots\ldots\ldots (42)$$

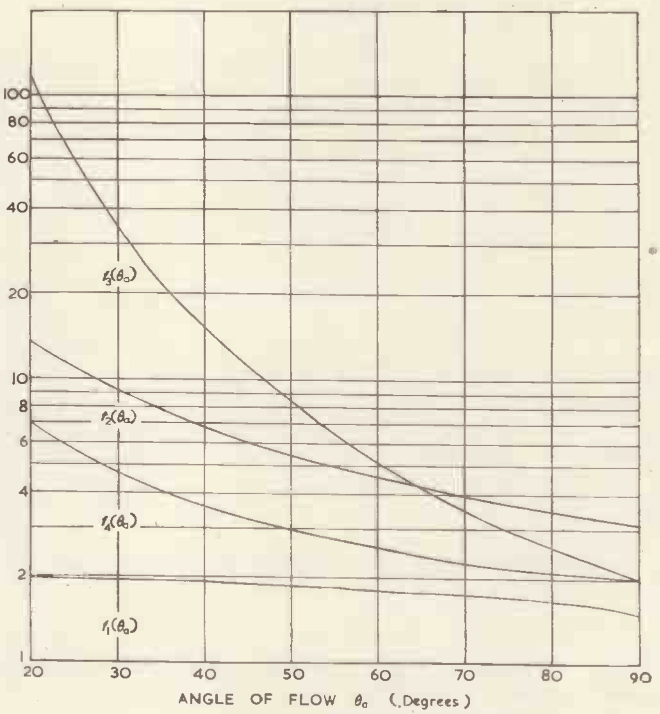


Fig. 3. Functions of θ_a for oscillator

The power dissipated at the grid electrode therefore becomes:

$$P_g = I_{g(dc)} \, V_g - I_{g(dc)}^2 \, R_g \quad \ldots\ldots\ldots\ldots (43)$$

$$= I_{g(dc)} \, v_{g(max)} \quad \ldots\ldots\ldots\ldots\ldots\ldots (44)$$

As the crest of the cycle is reached the anode voltage passes through its minimum value and causes the grid current to increase rapidly with the result that the grid current pulse is sharply peaked. The pulse shape approximates to a sine-squared function and accordingly the relation between the alternating and direct current components can be expressed as follows:

$$i_g / I_{g(dc)} = \frac{(V_g \cos \beta - V_B)^2}{1/\pi \int_0^{\theta_g} (V_g \cos \beta - V_B)^2 \, d\beta} \quad \ldots\ldots (45)$$

from which follows that as $V_B = V_g \cos \theta_g$ and $\cos \beta$ of the numerator equals unity at peak value:

$$i_{g(pk)} / I_{g(dc)} = \frac{\pi(1 - \cos \theta_g)^2}{\theta_g / 2 - 3/4 \sin 2\theta_g + \theta_g \cos^2 \theta_g} \quad (46)$$

Relations Governing the Anode and Grid Operating Angles

For a given valve the critical factors which govern the

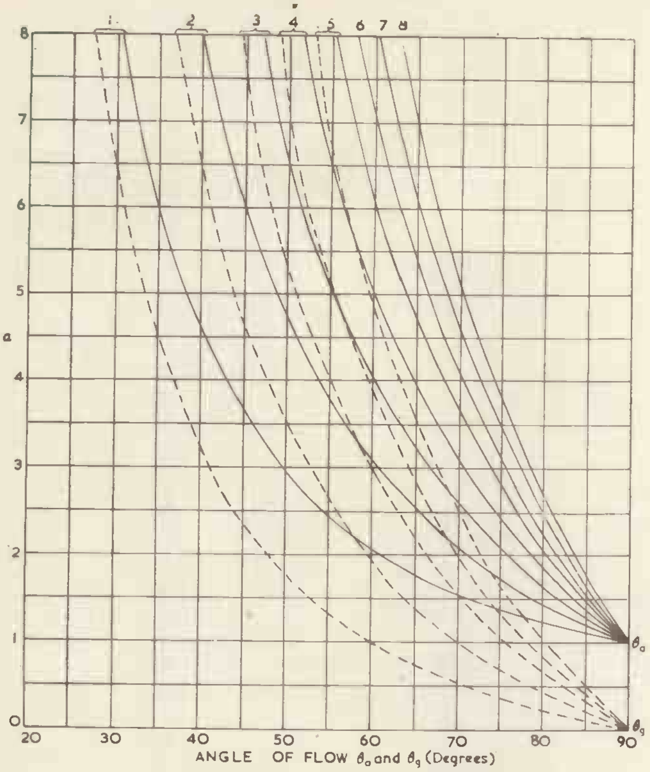


Fig. 4. Curves showing relation between α and γ for θ_a and θ_g when the value of $\frac{v_{a(min)}}{V_o}$ (τ) is taken as 0·1

anode operating angle $2\theta_a$ are the ratios $\mu V_B/V_o$, $\mu v_{g(max)}/V_o$ and $v_{a(min)}/V_o$ (τ). Calling the values of $\mu V_B/V_o$ and $\mu v_{g(max)}/V_o$, α and γ respectively, then from equations (21) and (22):

$$\theta_a = \cos^{-1}\frac{\alpha - 1}{\alpha + \gamma + \tau - 1} \quad \ldots\ldots\ldots\ldots (47)$$

Likewise for the grid operating angle:

$$\theta_g = \cos^{-1}\frac{\alpha}{\alpha + \gamma} \quad \ldots\ldots\ldots\ldots (48)$$

In the graph shown in Fig. 4 which has been prepared from these expressions the value of $v_{a(min)}/V_o$ has been taken as 0·1 and thus for a fixed value of $v_{a(min)}$, the grid bias voltage V_B, the grid exciting voltage V_g and the grid operating angle θ_g may be determined for a given anode operating angle θ_a once the maximum grid voltage $v_{g(max)}$, as chosen from the characteristic curves for a particular valve, has been established. Knowing the peak grid current drawn under the conditions specified, the direct current may be obtained from equation (46) and thus the value of the grid resistance required to produce the correct bias voltage may be determined. It can be seen that for a given duration of anode current flow there will be a limiting point beyond which the values of α and γ becomes unduly large, resulting in the power dissipated in the grid circuit becoming excessive. Usually α varies between 1·5 and 4 and γ between 2 and 8, the high values being demanded by valves of large impedance.

(To be continued)

A Single Cycle Timer for Small Spot Welders

By G. O. Crowther*, B.Sc., A.C.G.I., and L. H. Light*, B.Sc., A.Inst.P.

An improvement in the quality of spot welds in small components results if the welding time is reduced, thereby minimizing the region adjacent to the weld in which there is a risk of damage by over-heating or oxidation. The single-cycle timing circuit described was originally designed for use with existing welding equipment in valve manufacture, but can be adapted for other applications. Suggestions for modifying the circuit as a multi-cycle timer are also given.

THE quality of resistance welding may be considerably improved by reducing the welding time to as short a period as is consistent with the peak power handling capacity of the welding machine. In these circumstances, the heat, which is generated mainly at the spot where the surfaces to be joined meet, cannot spread appreciably from that region so that the area over which the metal is softened is minimized and the risk of burning or oxidation is almost eliminated. The circuit described in this article was developed for use with existing welders employed in the assembly of thermionic valves. Apart from one fundamental limitation, namely that the circuit is not suitable for welds to copper, the performance on a variety of operations was satisfactory and the expected improvements in the finished work were realized. The timer is considered suitable for any low-power welder operating from 200V to 250V a.c. mains and requiring a peak current within the rating of the thyratrons used, i.e. 40A for thyratrons type MT57 (XG1-2500).

* *The Mullard Radio Valve Co. Ltd.*

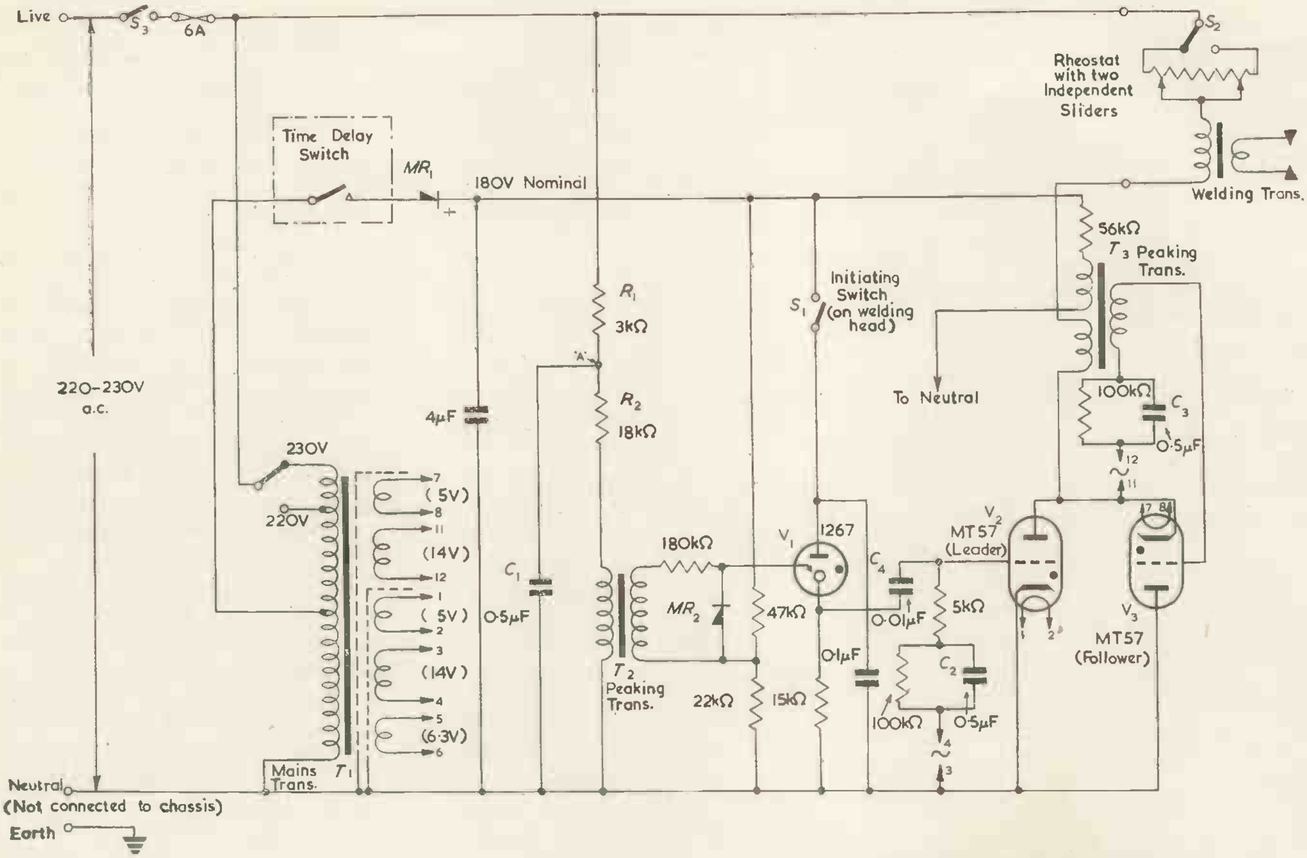
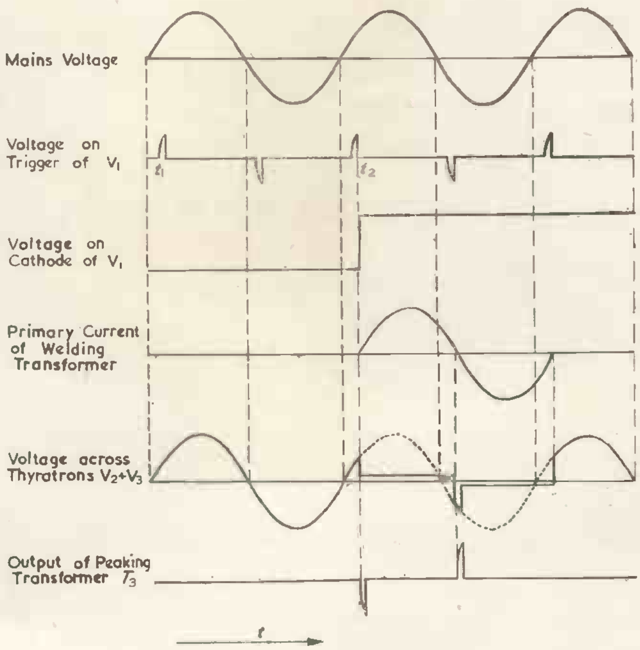


Fig. 1. The single cycle welding timer

Fig. 2. Waveforms in the single cycle timer

The sequence of events shown here ensues if closure of switch S_1 occurs between t_1 and t_2. It may open any time after t_2.



Design Considerations

Experiments carried out during the development of this timer indicated that a welding period of 20msec is optimum. This time proves short enough to realize most of the advantages of short-period welding without making necessary the use of larger and more expensive welding transformers. It also allows the use of very simple and reliable circuits.

In single-cycle welding it is essential to commence and finish the weld at the same point in the mains cycle, thus avoiding excessive variations of welding power, large switching surges, and saturation effects in the welding transformer. This demands much more accurate switching than is possible with mechanical relays. In the timer under review, switching is performed by two type MT57 thyratrons connected in inverse-parallel and in series with the main welding transformer. A cold-cathode trigger tube type 1267 (Z300T) is employed to fire one of the thyratrons which conducts for a half cycle. The second thyratron is arranged to conduct for the following half cycle.

Circuit

The circuit diagram of the timer is reproduced in Fig. 1 and the various waveforms are shown in Fig. 2. A weld is initiated when the operator depresses the pedal thus closing switch S_1, which connects the supply to the anode of the trigger-tube V_1. This tube remains non-conducting, however, until a timing pulse from the peaking transformer T_2 appears on the trigger electrode. When V_1 conducts, the potential at its cathode rises suddenly to $(V_{ht}\text{-}V_{burn})$, and thus fires the first thyratron, V_2, via C_4. When V_2 extinguishes at the end of the first half cycle, an impulse is generated by the "follow-on" peaking transformer T_3, firing V_3 which therefore conducts during the second half cycle.

Since V_1 remains conducting until the operator releases the pedal, thus opening switch S_1 and disconnecting the anode supply to the tube, the timing cycle cannot be repeated until the pedal is again depressed.

Peaking Transformers

The timing pulses on the grid of V_1 are obtained from the peaking transformer T_2, the primary of which is connected in series with resistor R_2 to a voltage derived from the mains. As indicated in Fig. 3, the alternating current I_P in the primary circuit is sufficiently large to saturate the core of the transformer for a considerable portion of the cycle. A voltage will appear across the secondary only during the short time when the core is unsaturated while the primary current is passing through zero. Because the circuit is almost entirely resistive ($R \gg \omega L_{unsat}$), the primary current will be in phase

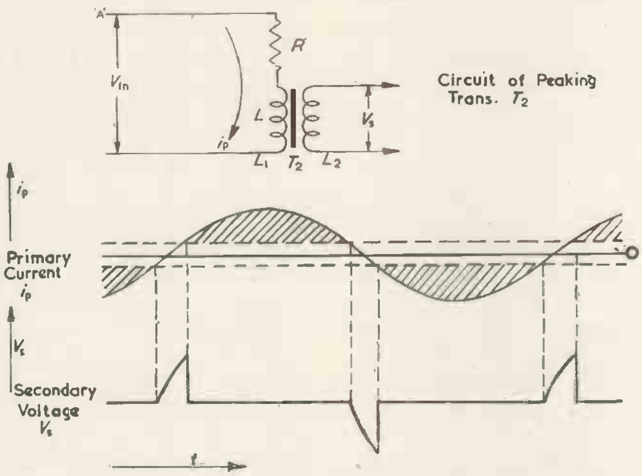


Fig. 3. Waveforms in the circuit of the peaking transformer T_2

In practice the core is saturated for a much greater part of the cycle than has been shown here. The actual duration of the secondary pulse is therefore much shorter.

with the voltage at the point A in Fig. 1. The phase-shift circuit R_1C_1 causes the voltage at A to lag 30° to 40° behind the mains so that the positive pulse produced in the secondary of the peaking transformer does not occur until the anode of V_2 is positive. The rectifier MR_2 suppresses the negative output pulses from T_2 so that they cannot interfere with the operation of V_1.

The second thyratron V_3 is fired by a pulse delivered by a similar transformer, T_3, the primary of which is connected in series with the primary of the main welding transformer. A positive pulse appears at the grid of V_3 as the current in V_2 falls to zero. Since, as shown on the left in Fig. 4, only half the range of primary current over which the core is unsaturated is used, the output pulse for a given turns ratio is less than if the full range ($-I_{sat}$ to $+I_{sat}$) were used. To obtain more efficient operation, therefore, a third winding has been added through which a steady direct current is passed. This current shifts the range of primary current over which the core is unsaturated from ($-I_{sat}$ to $+I_{sat}$) to (0 to $2I_{sat}$) so that the complete range of current over which the core is unsaturated is, in fact, employed, as shown on the right in Fig. 4.

As will be seen from Fig. 2, which shows the waveforms in the main welding circuit, the current in the primary of the main welding transformer lags behind the applied voltage since the circuit contains an inductive component. The anode-cathode voltage of V_3 will, however, remain negative until the first thyratron V_2 ceases to conduct, when it suddenly goes positive to a value equal to the instantaneous value of the mains voltage. At the same instant the large positive pulse from T_3 appears on the grid of V_3 and fires the tube.

Thyratron Bias

The two thyratrons are biased by alternating voltages

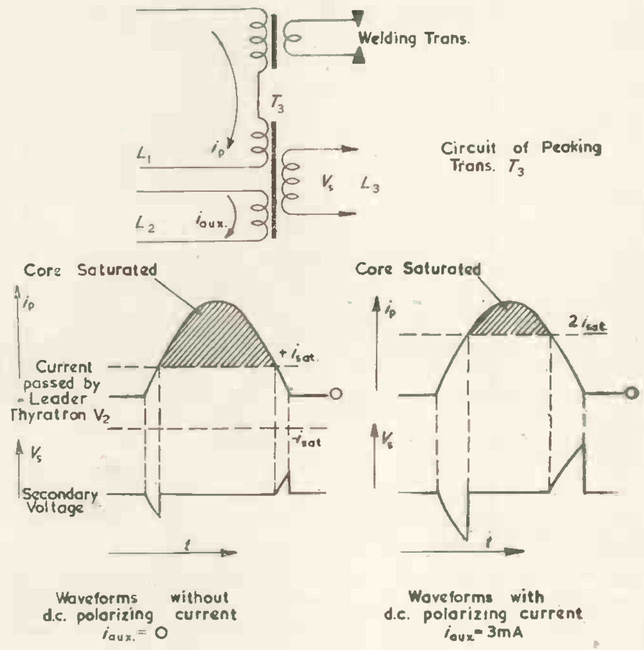


Fig. 4. Waveforms in the circuit of the peaking transformer T_3

in anti-phase with their respective anode voltages and obtained from separate windings of the filament transformer T_1. The grid current which flows during the negative half-cycle of anode voltage charges capacitors C_2 and C_3 in such a sense that a negative bias is superimposed on the alternating grid voltage sufficient to prevent the thyratrons from conducting during the succeeding half cycle.

Delay Circuit

A pre-heating period must be arranged before the thyratrons are allowed to operate. A delay of six to eight minutes, which is adequate for the MT57 mercury vapour valve in view of the very low duty cycle in this application, can be provided either by thermal delay mechanisms or by electronic means, such as the "boot-strap" delay circuit[1]. This should be connected in such a way that the anode circuit of V_1 is held open until the pre-heating period has elapsed, thus rendering premature triggering of the thyratrons impossible.

Input Voltage Range

The electrical design of the circuit has been calculated so as to ensure reliable single-cycle timing under all conditions likely to be encountered in normal practice. In particular, the circuit performance is not affected by mains voltage fluctuations even in excess of +10 or —15 per cent from the nominal value, which in the prototype equipment is 220 or 230 volts. When such large mains variations are expected to occur frequently, however, a constant voltage transformer should be used for the thyratron heater supply. Generous margins have been allowed for changes of valve characteristics over life and for component tolerances.

Power Handling Capacity

Owing to their very brief and infrequent conducting periods, the MT57 thyratrons employed in the timer can be run at their "igniter service" peak current rating of 40A. This was found to be ample for all welds encountered in receiving valve assembly. Larger welding power can also be obtained by using a higher input voltage. If this is done, say by connecting the welder between two phases of a three-phase supply, it is necessary to increase the a.c. grid bias voltage of V_2 and V_3.

Installation

Weld-initiating Switch

The switch operated by the pedal for initiating the weld must be of a type which makes positive contact without excessive bounce, otherwise there is a risk of occasional double-cycle welds. Micro-switches have been found to be satisfactory, and should preferably be installed on the welding head in such a position that they close when a definite force is applied to the work, irrespective of the adjustment of the welding electrodes.

Heat Control

Since the welding time is fixed, the only way of varying the energy input to the weld is by adjusting the welding power. This can be done most conveniently by arranging a variable resistor in the primary circuit of the welding transformer. Fig. 1 shows an arrangement in which two preset values of welding energy are obtained by means of a changeover switch which selects part of a single rheostat having two taps.

Interference

The apparatus is insensitive to all forms of mains interference. This property is most fully realized when the live and neutral leads of the mains supply are connected as shown in Fig. 1, and this method of connexion should therefore be adopted whenever possible.

Forging Pressure

Forging pressures higher than normal are required in short-period welding. This pressure must remain effective while the material "gives" during the current flow. The electrode tips must thus be capable of being accelerated quickly, making low inertia of the moving parts or sufficient compliance in the electrode system highly desirable.

Performance

The prototype timer was fitted to standard welding equipment modified only to give an increase in forging pressure. Prolonged trials in various departments of the valve factory have shown that very satisfactory results can be obtained with the timer. The quality of welds produced is high: strength, consistency, appearance and freedom from oxidation are all substantially better than with multi-cycle timing. The improvement obtained has proved particularly valuable in the welding of difficult materials such as molybdenum. The factory has now been largely re-equipped with welders incorporating the single-cycle timer. The timers have proved reliable and a very high standard of welding has been achieved under production conditions.

The fact that a pre-heating time of approximately six minutes is required before commencing welding is a minor disadvantage which, however, has not proved a serious handicap in practice. A more serious disadvantage is that single-cycle welding cannot be used for welding copper to other metals. The reason is that insufficient time is available for the heat to spread to the low-resistivity copper from the other metal where most of the heat is generated. This is a fundamental limitation which imposes some restriction on the use of the process.

Adaptation for Multi-cycle Welding

It is possible to design multi-cycle timers using the same operating principles as the single-cycle timer described. The cold-cathode tube V_1 in the single-cycle timer is, in effect, a gate which allows only one of a train of trigger pulses to produce an output from the two thyratrons each time the weld-initiating switch is closed. This constitutes a single counting action—a count of one.

The principle of using digital methods instead of analogue methods can also be used with advantage in multi-cycle timers, particularly in those which employ thyratrons to switch the welding current. Simple counting circuits, which pass the requisite number of pulses to the leader thyratron each time the initiating switch is closed, can be designed round one of the valves recently developed for the purpose, such as the cold-cathode ring counters or the EIT hot-cathode ribbon-beam counting tube[2]. Such timers have the advantage of synchronous operation and absolute accuracy which is often difficult to achieve in other timing methods.

Acknowledgments

The authors are indebted to Mr. Harley Carter for valuable assistance in the preparation of the manuscript, and to the Directors of the Mullard Radio Valve Co. Ltd. for permission to publish this article.

Appendix

Details of Peaking Transformers T_2 and T_3.

Peaking Transformer T_2

L_1 5000 turns 47 s.w.g.

L_2 10 000 turns 47 s.w.g.

Peaking Transformer T_3

L_1 100 turns 20 s.w.g.

L_2 2000 turns 47 s.w.g.

L_3 3000 turns 47 s.w.g.

Core of T_2 and T_3: 6 Mumetal laminations interleaved U-shaped stamping No. 30, thickness 0·4mm.

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A Portable Restriking Voltage Display Unit

By J. W. Armitage*, Dipl.Eng., A.M.I.E.E.

This article describes a method of displaying on the screen of a cathode-ray oscilloscope, employing a stationary film camera and a continuously running time-base, the whole history of the transient developed during separation of switch contacts. Although the instrument was made for the investigation of restriking voltages appearing across circuit breaker contacts on separation and for measuring the surge voltages during fuse short circuit-testing, it can be used for studying relays, contactors, etc. The time-base is continuously running and the rapid flyback results in the transient being divided into a number of small sections which are superimposed on the same portion of the c.r.o screen. Thus the whole history of the transient is recorded on a film approximately 1in square. Some records taken during actual circuit breaker and fuse tests are given.

THIS equipment has many advantages over other and older types of equipment used for this work which were of the continuously evacuated type and used rotating drum cameras. The drum with film rotating in vacuo. The instrument described uses a normal c.r.o. recording camera incorporating a bloomed *f*/1, Wray, c.r.o. copying lens. Using the same technique it is also possible to record three phases simultaneously by using a single tube unit in each phase, or current and voltage can be investigated by a twin tube unit.

The unit here described records on a stationary film the transient voltage and subsequent 50c/s open-circuit voltage across the breaker by means of a sealed-off cathode-ray tube. In order to obtain sufficient length of trace to show the details of the front and subsequent high frequency portion of the transient with sufficient clarity a continuously running time-base, with rapid flyback, was employed. Thus the recorded voltage is divided up into a number of short time lengths, all superimposed on the same portion of the fluorescent screen of a cathode-ray tube. To facilitate time calibration of the records the time-base is synchronized to the calibration oscillator. At first sight it seems that the records are somewhat crowded, but the transient is quite clearly recorded and all the necessary information about the transient easily determined without reference to divider characteristics etc. Direct comparison between transient peak and 50c/s wave peak gives the transient amplitude, frequency of transient again is obtained by direct comparison with calibration oscillator as is the front of the transient, also duration of transient can be determined, in fact the whole history of the transient is contained on the 1in square film.

The small size of the finished record is a great advantage since filing space can be drastically reduced. The instrument also has the merit of being transportable and is always ready for service.

Since the instrument employs a continuously running time sweep provision is made for brightening the cathode-ray beam in, for a brief period before the circuit breaker contacts open, or the fuse blows, and suppressing it after the transient has passed. The total length of time that the beam is brightened is variable and hence the total length of the complete record can be varied to suit the test conditions. An internal trip circuit is included so that the calibration wave can be shown on the record. Also to avoid confusion of the main trace, when the calibration wave is switched on, the tube trace is displaced in the Y direction, and by synchronizing the time-base, at all times, to the calibration oscillator, time calibration is simplified since this ensures that the calibration wave appears as a fine trace.

Description

The equipment consists of (a) a cathode-ray tube, (b) 500V stabilized power supply, (c) high frequency fed d.c. set for supplying the 6kV beam accelerating voltage, (d) a continuously running time sweep operating on the Miller principle, (e) calibrating oscillator giving 500c/s and 10kc/s, (f) auxiliary circuits including tripping, beam modulation and tube controls, etc.

The equipment was housed in two steel cabinets, the power supplies, stabilizer and h.f., d.c. unit being built in the larger one and the smaller one containing the cathode-ray tube on its upper deck and the time sweep, modulator and timing generator on its lower deck.

Circuit Operation

The circuit diagram is shown in Figs. 1 (a), (b) and (c) for the power pack cabinet and in Figs. 1 (d), (e) and (f) for the c.r.o. cabinet. The +500V d.c. supply (Fig. 1 (a)) is obtained from the transformer T_1 and full wave rectifier V_1 and supplies the other circuits via the series fed stabilizer V_3 and V_4, the reference voltage being that developed across the neon V_5. A further rectifier V_2 supplies the 500V negative d.c. for biasing and for providing a negative supply for the push-pull shift for the X and Y plates of the cathode-ray tube.

The cathode-ray tube is supplied from the h.f. unit (Fig. 1 (b)) which delivers about 6kV. The supply to the various anodes is taken from the divider chain R_{52} to R_{56} (Fig. 1 (e)). The heater supply for the tube was originally taken from a third winding on the r.f. transformer T_3 but this did not prove too satisfactory as modulation of the beam occurred. The heater supply was run from a 50c/s transformer and the fault was eliminated. Positioning of the spot on the screen is accomplished by the ganged potentiometers R_{61} and R_{62}, R_{63} and R_{64} (Fig. 1 (e)). To prevent the rectifier V_8 from having voltage on its anode

* *Metropolitan-Vickers Electrical Co. Ltd.*

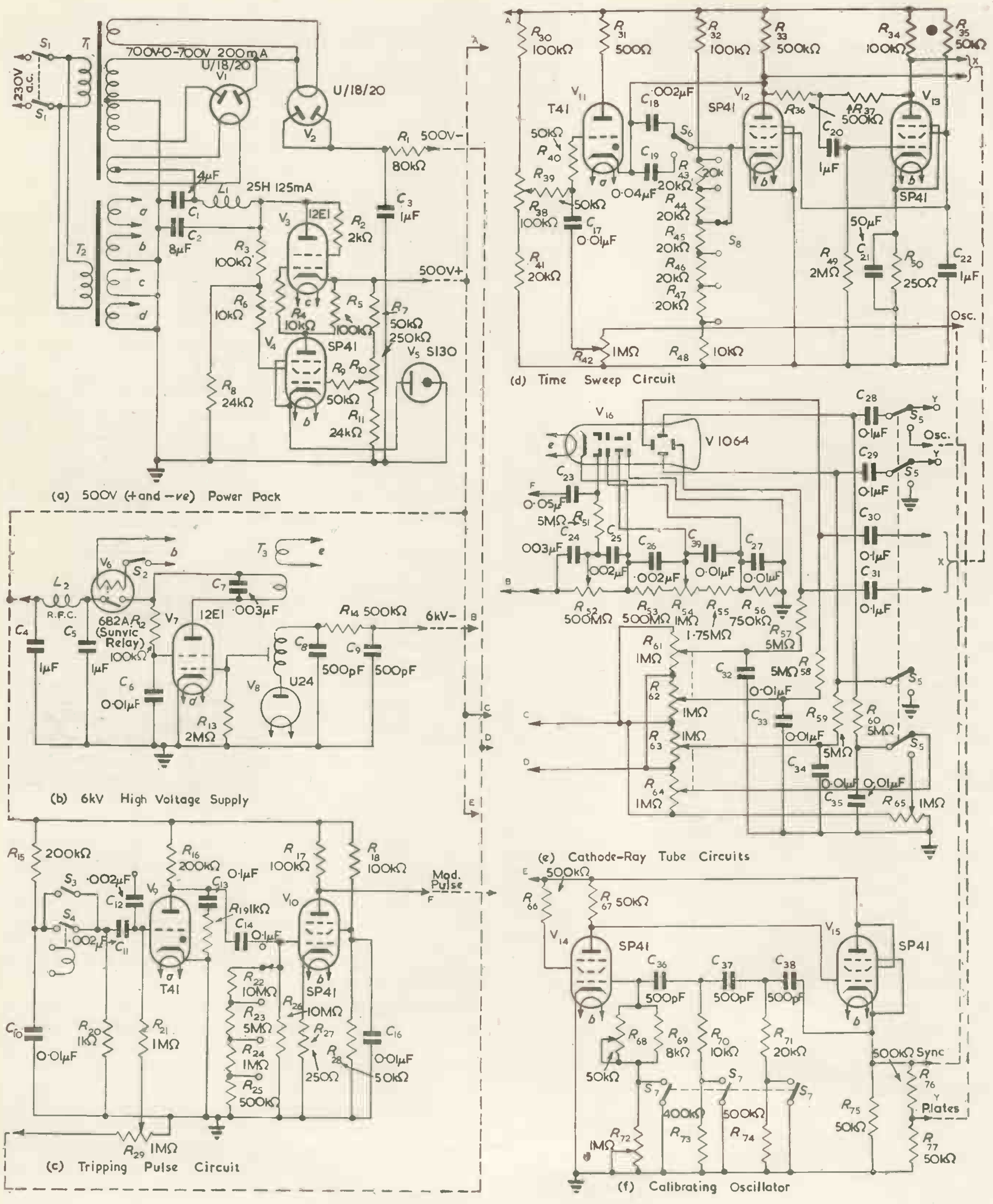

Fig. 1(a-f). The transient cathode-ray oscillograph

before its filament is heated a "Sunvic" time delay V_6 is fitted to the 500V supply to the h.f. unit.

The time sweep (Fig. 1 (d)) operates on the "Miller" principle, the capacitor C_{18} or C_{19} being charged through the resistor R_{31} by thyratron V_{11} and discharged through the pentode V_{12} from anode to grid. The flyback occurs during the charging period and the sweep proper is controlled by the discharge through V_{12}. Control of the sweep speed is by means of the resistance chain R_{32} to R_{48}. The sweep amplitude is controlled by R_{38} by limiting the voltage to which capacitor C_{18} or C_{19} can be charged. Since a symmetrical sweep was required it was necessary to fit a phase reversing circuit, the voltage appearing at the anode of V_{12} being applied to the grid of V_{13} via resistor R_{36} and C_{20}. The anode of V_{13} thus reproduces the same waveshape as appears on the anode of V_{12} but of opposite polarity. The time sweep plates are thus fed from the anodes of V_{12} and V_{13} via the coupling capacitors C_{30} and C_{31} respectively. With the phase reversing circuits, sweep amplitudes of from zero to about 900V can be obtained from a supply voltage of 500V d.c., the maximum voltage change appearing at the anodes of V_{12} and V_{13} being about 450V. The time sweep generator is synchronized to the calibration oscillator by injecting a portion of the oscillator voltage which appears across potentiometer R_{42} into the grid of thyratron V_{11} by means of capacitor C_{17} and R_{40}.

The timing calibration is provided by means of a simple *RC* oscillator (Fig. 1 (f)). The choice of an *RC* oscillator was governed by the frequencies demanded, namely, 500c/s and 10kc/s. An *LC* oscillator to provide 500c/s would be physically large. It will be noted that the changeover switch from 500c/s to 10kc/s is ganged to the changeover switch for C_{18} or C_{19}. Thus the 500c/s calibration wave appears only on the slow sweep speeds which range from about 5msec to 100msec and the 10kc/s on the faster range of from 250μsec to 5msec. Also ganged with this changeover switch is a further switch for earthing one Y plate and adding a fixed bias to the other so that the calibration wave does not interfere with the main trace.

Since the 125V d.c. used for tripping the circuit breakers under test, contained considerable ripple which was sufficient to trip off the thyratron V_9, the post office relay S_4 was substituted. The 125V supply operates this relay, applying a positive pulse to the grid of V_9 via C_{10} and R_{20} and this in turn makes pentode V_{10} conductive and applies a positive pulse to the grid of the cathode-ray tube via capacitor C_{23}. The internal trip button S_3, being connected across the P.O. relay contacts, performs the same operation and is used to apply the calibrating wave. The grid of V_{12} is also brought out to a terminal on the front panel marked "External Trip" to synchronize modulation with other external equipment. Control of the length of modulating pulse is obtained by the resistors R_{22} to R_{26} and capacitor C_{14}. The steps on the resistor chain are made to correspond to a modulation time of approximately one to four cycles of the 50c/s supply. This is necessary since the precise time of opening of a circuit breaker may not be known and if the spot remains illuminated too long the maze of 50c/s lines appearing on the screen would render the transient illegible.

Results Obtained

Fig. 2 shows the record of transient recovering voltage obtained during the short-circuit testing of an air blast breaker. In the case of Fig. 2 (a) the total duration of the modulation is about 25msec, i.e. the total length of trace is 25msec but subdivisions are each of 1msec. The calibration oscillator is 10kc/s. Thus the following information can be deduced for the record. Maximum voltage rise of transient equals 1·4 times peak alternating voltage, frequency of resulting oscillations about 9kc/s. Total duration of transient 1·5msec, time of rise of front of transient 55μsec.

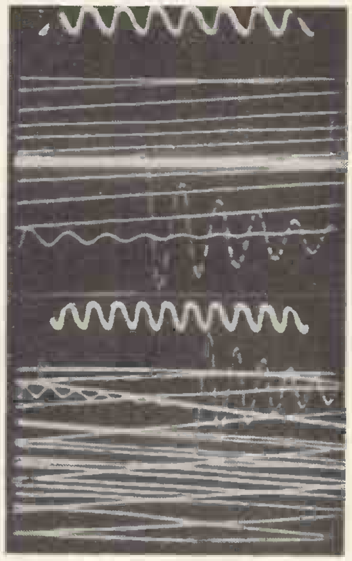

Fig. 2. Oscillograms taken during testing of air blast breakers

(*a*) *Total sweep duration = 25msec. Duration of each forward trace = 1msec. Calibration oscillator = 10kc/s. Peak of transient = 1·4 × a.c. peak. Transient frequency = 9kc/s. Rise time of transient front = 55μsec.*

(*b*) *Total sweep duration = 40msec. Duration of forward trace = 1·2msec. Calibration oscillator = 10kc/s. Peak transient = 1·5 × a.c. peak. Transient frequency = 10kc/s. Rise time of transient = 50μsec.*

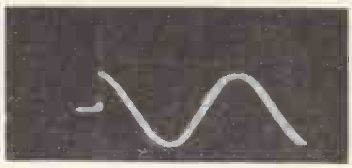

Fig. 3. Oscillograms taken during fuse testing

Peak voltage attained during fuse blowing = 2·5 × a.c. peak.

In the case of Fig. 2 (b) the following information is available: total duration of sweep 40msec, sweep length 12msec. Calibration oscillator is 10kc/s, which gives a transient frequency of about 10kc/s. Peak transient voltage is 1·5 times peak alternating voltage.

Fig. 3 shows the record obtained of the transient recovery voltage produced in the short-circuit test on a high voltage fuse. In this case a detailed knowledge of the waveshape of the recovery voltage was not required but what was of interest was the peak value of this voltage. Consequently the time sweep and beam modulation controls were so adjusted as to give a record showing a single sweep in the forward direction of duration equal to 1½ cycles of the 50c/s voltage. Although, as a result, the record of the transient voltage is compressed on the time axis, its peak value is clearly indicated and can be determined by comparison with the amplitude of the 50c/s voltage shown.

The recommended test (see B.S.88*) for high voltage fuses requires that a sphere gap, set at a suitable voltage,

* *Also A.S.T.A. Rules No. 8, Governing the Short Circuit Testing of H.V. Fuses.*

should be connected in parallel with the fuse during the short-circuit test and that any recovery voltage developed when the fuse blows should be insufficient to flash over the sphere gap. This test, while reasonably simple to perform, merely shows whether the fuse has passed or failed, but does not indicate by what margin either of these results is achieved. The use of the display unit, however, in the manner indicated gives a positive measurement of the peak value of the recovery voltage and this can be compared with the stated peak voltage in the specification to decide whether the fuse has passed the test.

In practice it is found equally simple to use the display unit as to use the sphere gap and in the case of high voltage fuses it is always much more convenient to use the former device. The display unit, moreover, has the advantage of giving more information and in the form of a permanent record on a photographic film.

Conclusions

The restriking voltage display unit using a sealed-off cathode-ray tube has satisfactorily performed the function for which it was designed, namely, of giving a record, lasting as desired from 1 to 5 cycles of a 50c/s wave, in the form of a series of short sweeps of a duration which can be varied from 250μsec to 1msec.

It is capable, therefore, of recording such long-time phenomena as the transient recovery voltages of circuit breakers with a writing speed high enough to show details of waveshape and amplitude of the high frequency portion of the wave. The recorded trace can be photographed on 35mm film, thus giving a compact negative requiring little storage space.

The display unit offers a convenient and more informative alternative to the sphere gap as used in the short-circuit testing of fuses and its use in this test is recommended.

Acknowledgments

The author wishes to thank Dr. Willis Jackson, F.R.S., Director of Research and Education, and Mr. B. G. Churcher, M.Sc., M.I.E.E., Manager of the Research Department, Metropolitan-Vickers Electrical Co. Ltd., for permission to publish this article.

A Technique for Non-Linear Function Generation

By P. N. Nikiforuk*

A method is described for converting an element that takes the n^{th} root (or n^{th} power) of a signal to one that performs the inverse operation. A particular non-linear element is described and experimental results given.

OCCASIONS arise when it is desirable to convert a non-linear element whose output is proportional to the n^{th} root (or n^{th} power) of the input signal to one whose output is proportional to the n^{th} power (or n^{th} root) of the input. This is desirable for example in certain types of non-linear servomechanisms where it is necessary to have part

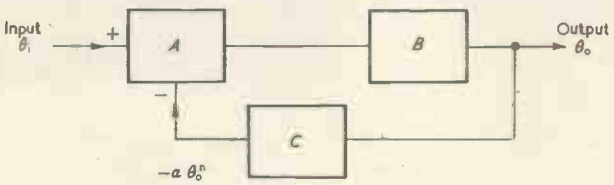

Fig. 1. Block diagram of non-linear circuit

of the control signal proportional to either the square root of the error signal or to the square of the velocity signal.

The method described in this article accomplishes this merely by using two pentode amplifiers in addition to the given non-linear element.

Principle of Operation

The n^{th} root (or n^{th} power) of a signal may be extracted by using a three element closed-loop system in such a manner that the desired output is the input to a non-linear element which performs the inverse operation.

In Fig. 1 the unit C is the given non-linear element and gives an output proportional to the n^{th} power of its input (θ_o). The unit A is an adding unit and combines the input signal and feedback signal to give an output equal to:

$$-(\theta_i - \alpha\theta_o^n) \quad \ldots\ldots (1)$$

The unit B is a high gain amplifier with a gain of G and gives an output equal to:

$$G(\theta_i - \alpha\theta_o^n) \quad \ldots\ldots (2)$$

This is equivalent to θ_o.

Hence:

$$\theta_o = G(\theta_i - \alpha\theta_o^n) \quad \ldots\ldots (3)$$

If $G \gg 1$ and $\alpha \simeq 1$, then:

$$\theta_o = (\theta_i/\alpha)^{1/n} \quad \ldots\ldots (4)$$

which is the desired result.

The Practical Circuit

The non-linear system shown in Fig. 2 has been used for some time in the Servo-mechanisms Laboratory at Manchester University to obtain an output signal that is proportional to the sign of the input signal multiplied by either the square root or the square of the modulus of the input signal. The system consists of three separate units, an adding element of the anode-follower type with a gain of unity, an amplifier with a gain of about 75 and a non-linear element whose output is proportional to the sign of its input (θ_o) multiplied by the square of the modulus of θ_o. In the arrangement of this circuit, which is basically a feedback amplifier of variable gain, the input signal is applied across two sets of diodes. Each diode is biased off NV_c volts more than the preceding diode where N is an integer. When the modulus of the input is less than V_c the gain is $-R_f/R_i$. As the input signal increases either positively or negatively the upper or the lower set of diode rectifiers becomes operative, the effective input resistance R_i' becomes small and consequently the overall gain $-R_f/R_i'$ large.

The curve relating input and output is approximated by segmented straight lines and the resulting function is parabolic at the points

$$V_c \ldots (M - N_2)V_c, (M - N_1)V_c, MV_c$$

* *Servo-mechanisms Laboratory, University of Manchester.*

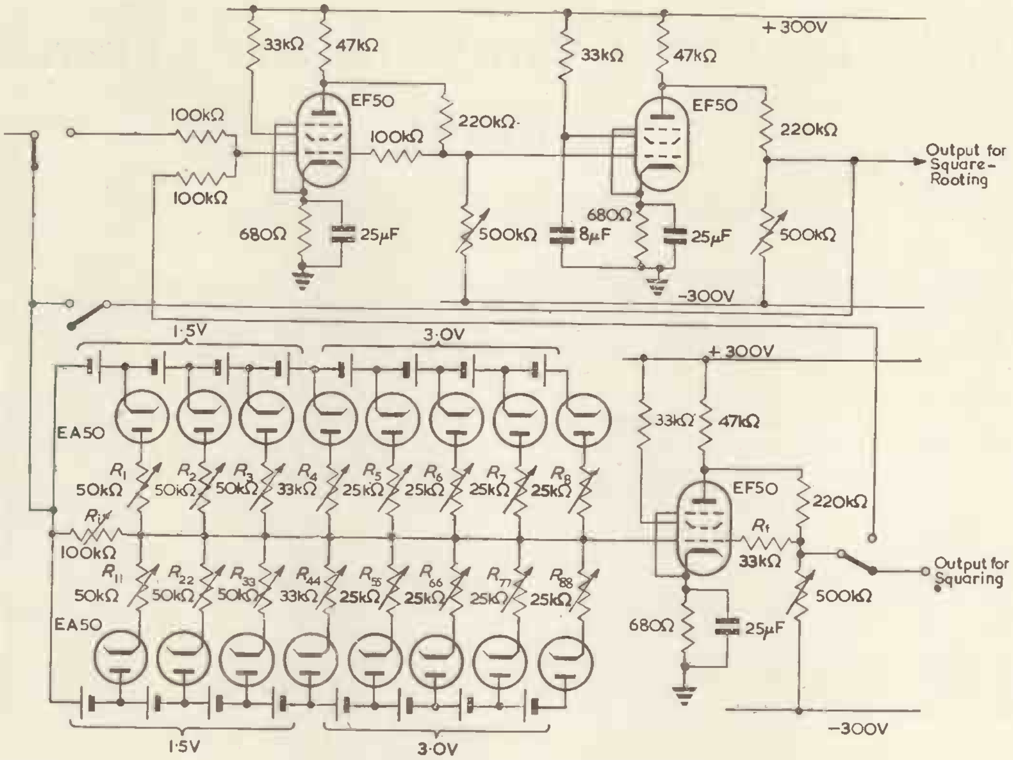

Fig. 2. Circuit for squaring or square-rooting
For squaring switches in the down position. For square-rooting switches in the up position.

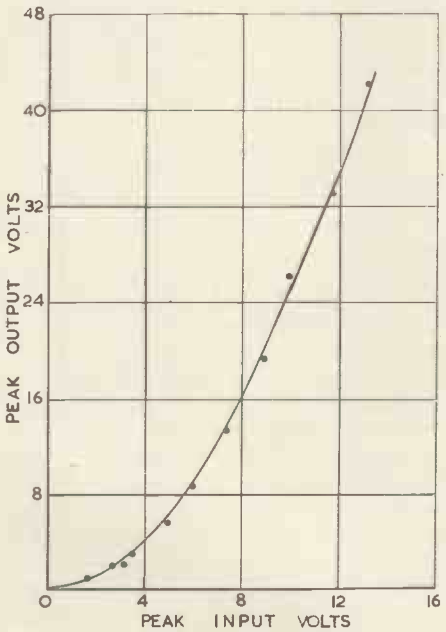

Fig. 3. Response of squaring circuit
Required function $Y = 0{\cdot}24\,X^2$.
Theoretical curve— Experimental points ·

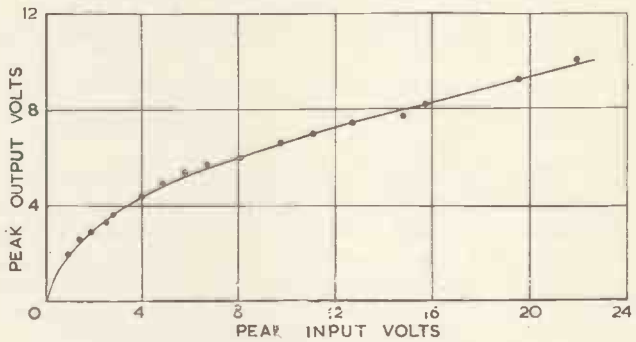

Fig. 4. Response of the square-rooting circuit
Required function $Y = 2{\cdot}1\sqrt{X}$.

if the slope over the successive intervals of $(M - N)V_0$ volts increases in the ratio:

$$\frac{M^2 - (M - N_1)^2}{(M - N_1)^2 - (M - N_2)^2} \cdot \frac{N_2 - N_1}{N_1}$$

where M, N_1, N_2 are integers, and $M, N_1, N_2 > 0$.

$R_1 \ldots R_8$ and $R_{11} \ldots R_{88}$ were calculated approximately. Variable resistors were then inserted in series with the calculated values and adjusted to give the desired characteristic.

Experimental Results

Experimental results together with calculated values are shown in Figs. 3 and 4. There is good agreement between experimental and theoretical values.

Conclusions

A simple method has been outlined for converting an element whose output is proportion to the n^{th} root (or n^{th} power) of the input to one which performs the inverse operation. The advantage of this method is that only two pentodes and a few standard components are necessary. An experimental system utilizing this principle has been described and experimental results given. The experimental results verify the reliability of the operation.

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Analysis of the Common-Base Transistor Circuit

By Francis Oakes*, M.Inst.E., A.M.Brit.I.R.E.

A hybrid inverted π-network is chosen as the equivalent circuit for the grounded base transistor amplifier. This permits the derivation of simple equations for important circuit parameters, and provides useful concepts for understanding the interdependence of input and output circuits, stability criteria and other operating characteristics of this basic transistor circuit.

THE common base circuit is widely used in point contact transistor applications. Connected in this way, the transistor exhibits negative input and output resistance characteristics under certain conditions and lends itself to oscillator and switching applications. On the other hand, circuit components can be chosen so that the transistor functions as a stable amplifier.

A number of descriptions of the common base circuit have been published[1] containing *T*-derived equivalent circuits and, based on these, derivations of stability criteria, input and output resistance etc. Other treatments are based on matrix representation of the transistor circuit and the same expressions result from a purely algebraic approach. While such methods are quite suitable for deriving formulæ, they do not lend themselves readily for assessment of parameter influence on circuit behaviour, nor are they very conducive to understanding the operation of transistor circuits. It is the purpose of this article to present an alternative method, designed to characterize the behaviour of the circuit, and to aid in assessing the factors influencing the operation of transistor oscillators and amplifiers.

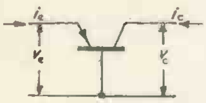


Fig. 1. Common-base transistor circuit

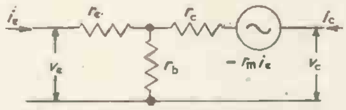


Fig. 2. T-derived equivalent circuit for the common-base transistor circuit

r_e = emitter resistance, r_c = collector resistance, r_b = base resistance, r_m = mutual resistance

With the base of the transistor grounded, and the input applied to the emitter, a circuit is obtained as shown in Fig. 1. It can be demonstrated that the action of a transistor, so far as its small signal behaviour is concerned, is characterized by an equivalent circuit consisting of three impedances (these are resistive at low frequencies) plus a constant voltage source as shown in Fig. 2. The arrows are used to define the positive direction of current flow. These directions are, of course, arbitrary, and it is merely a convention to define the direction of current flow into the network as positive.

Although this equivalent circuit is used quite widely in the literature, and was chosen as the basis of analysis in the references quoted, it is quite easy to see that the equivalent circuit shown in Fig. 2 is not a very practical one. For instance, α, the current gain, does not appear as a parameter, in spite of being one of the most important characteristics of the transistor. Furthermore, measurement of voltages, currents, admittances or transfer ratios for instance, will not yield the four parameters directly, and further calculations will be required. For example, the parameter r_e would have to be found by the measurement of the input resistance $r_e + r_b$, and measurement of the feedback resistance r_b, and subsequent subtraction of the latter from the former.

Thus, it is preferable for many applications to describe a transistor in terms of directly measurable parameters such as the input and output resistance (impedance or admittance), the current gain and the feedback ratio, instead of the four resistances of Fig. 2. Taking this into account, a more suitable equivalent circuit is gradually coming into use[2]. This circuit is the inverted π equivalent circuit. A version of this circuit, which is shown in Fig. 3, is used as the basis for analysis of the common base circuit. As can be seen from the illustration, it contains two resistances (which become reactive at higher frequencies) and two sources, one of them a constant voltage, the other a constant current generator.

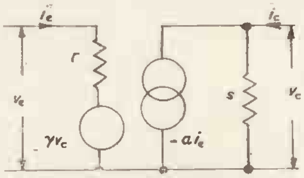


Fig. 3. Inverted π-derived equivalent circuit for the common-base transistor circuit

Typical values: $r = 100\Omega$, $s = 20k\Omega$, $\alpha = 2$, $\gamma = 0{\cdot}01$

For most junction transistors using grounded base and input emitter connexions, the input conductance is so great and the output resistance so high that the input voltage and the output current cannot be controlled easily. But even with point contact transistors, where they can be controlled, it is advantageous not to choose them as the controlled variables. The remaining variables, namely input current and output voltage, on the other hand, are very suitable to be chosen as controllable for test or design purposes. The equivalent circuit shown in Fig. 3 takes this into account and it is seen that the sources included are characterized by i_e and v_c, i.e. by those parameters which are easily controlled. Furthermore, the parameters used for this circuit, namely r, s, α, γ, are much more characteristic of transistor action than those of Fig. 2, and lend themselves readily for direct measurement.

It can be seen from Fig. 3, by inspection that:

- r—is the input resistance for shorted output.
- γ—is the ratio of input to output voltage for open-circuit input terminals.
- s—is the output resistance for open-circuit input terminals.
- α—is the negative ratio of output to input current for shorted output.

* *Ferguson Radio Corporation Ltd.*

The above statements define the parameters in the same way as does the equivalent circuit.

It is important to realize that in the output circuit, the constant current generator $-\alpha i_e$ is the "active" parameter. In the input circuit, the constant voltage generator γv_c merely represents a feedback voltage. The parameter r on the other hand, can become "active" in those transistors where it assumes a negative value of resistance.

The equivalent T and equivalent inverted π circuits and their parameters can be correlated by inspection of Figs. 2 and 3. Since they are equivalent to each other, the input and output voltages must be the same under equal external conditions. This offers easy means for finding the required correlations. For open-circuit output and unity input current the following equations are obtained:

$$v_c = r_b + r_m \quad \ldots \quad (1)$$

$$v_e = r_b + r_e \quad \ldots \quad (2)$$

but also:

$$v_c = \alpha s \quad \ldots \quad (3)$$

$$v_e = r + \alpha\gamma s \quad \ldots \quad (4)$$

Similarly, for open-circuit input and unity output current:

$$v_c = r_b + r_c \quad \ldots \quad (5)$$

$$v_e = r_b \quad \ldots \quad (6)$$

but also:

$$v_c = s \quad \ldots \quad (7)$$

$$v_e = \gamma s \quad \ldots \quad (8)$$

Equating (1) and (3), (2) and (4), (5) and (7), (6) and (8) four equations are obtained:

$$r_b + r_m = \alpha s \quad \ldots \quad (9)$$

$$r_b + r_e = r + \alpha\gamma s \quad \ldots \quad (10)$$

$$r_b + r_c = s \quad \ldots \quad (11)$$

$$r_b = \gamma s \quad \ldots \quad (12)$$

and from the last four equations, the following are obtained:

$$r_b = \gamma s \quad \ldots \quad (13)$$

$$r_c = (1 - \gamma)s \quad \ldots \quad (14)$$

$$r_m = (\alpha - \gamma)s \quad \ldots \quad (15)$$

$$r_e = r + (\alpha - 1)\gamma s \quad \ldots \quad (16)$$

$$r = r_e - r_b(\alpha - 1) = r_e - r_b\left(\frac{r_b + r_m}{r_b + r_c} - 1\right) \quad \ldots \quad (17)$$

$$\gamma = \frac{r_b}{r_b + r_c} \quad \ldots \quad (18)$$

$$\alpha = \frac{r_b + r_m}{r_b + r_c} = (r_b + r_m)/s \quad \ldots \quad (19)$$

$$s = r_b + r_c \quad \ldots \quad (20)$$

In this way, each set of four parameters can be calculated in terms of the other set of four parameters.

Equation (17) is of particular interest. It is well known that the main difference between junction and point contact transistors is that α cannot be greater than 1 for junction transistors. Thus, since $r = r_e - r_b\,(\alpha - 1)$, it follows that in a junction transistor r can never become negative, while for point contact transistors where α is of the order of about 2, r will become negative, especially where r_e is smaller than r_b.

An example is given below to illustrate the usefulness of the inverted-π circuit. The diagram is shown in Fig. 4(a) and the equivalent circuit in Fig. 4(b) gives values for the four parameters which are typical (for some types of point contact transistors). The transistor is fed from a constant voltage generator delivering 5mV in series with a generator resistance of 1 000Ω. The output is terminated by a resistance of 20 000Ω. By inspection of the emitter mesh

$$0{\cdot}005 = 900\ i_e + 0{\cdot}01\ v_c$$

by inspection of the collector mesh

$$v_c = \alpha i_e \times 10\,000 = 20\,000\ i_e$$

introducing this into the first equation

$$0{\cdot}005 = (900 + 0{\cdot}01 \times 20\,000)\ i_e$$

$$\left.\begin{aligned} i_e &= 4{\cdot}55\mu\text{A} \\ v_c &= 91\text{mV} \end{aligned}\right\}$$

dependent variables

$$A = 18 = \text{voltage amplification}$$

Thus it is seen that the collector voltage appears as 91mV and a voltage amplification of 18 is obtained. It must be stressed, of course, that as with all other such representations, this only holds for the small signal parameters at the particular operating point chosen. There is no difference in this respect between this and any other method such as use of matrices or the application of T derived equivalent circuits.

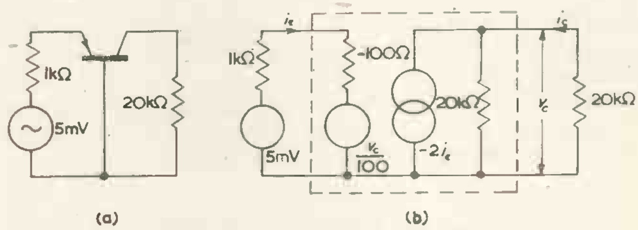


Fig. 4. Actual and equivalent circuit of transistor with generator and load

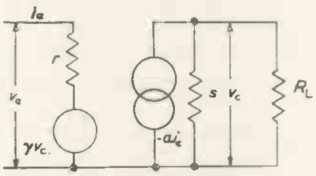


Fig. 5. Derivation of input resistance

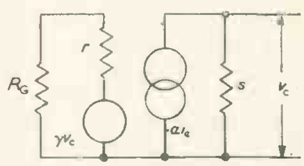


Fig. 6. Derivation of output resistance

Input and Output Resistances

The general behaviour, and in particular the stability or instability of a transistor circuit is characterized largely by the input and output resistances R_{in} and R_{out}. The equivalent circuit offers a ready means for deriving the input and output resistances and for assessing their dependences on internal, as well as external circuit parameters.

In the literature[1], input and output resistances are usually presented in the following form, which is not very descriptive:

$$R_{in} = r_e + r_b - \frac{r_b(r_b + r_m)}{R_L + r_c + r_b}$$

$$R_{out} = r_c + r_b - \frac{r_b(r_b + r_m)}{R_G + r_e + r_b}$$

Better expressions for input and output resistances can be derived quite easily from the circuit of Fig. 3. Thus the input resistance is found as the ratio v_e/i_e with a load resistance R_L terminating the output. This is shown in Fig. 5. The equations can be written down by inspection. From the output mesh one obtains:

$$v_c = +\alpha i_e\ R_L^* \quad \ldots \quad (24)$$

where:

$$R_L^* = R_L \| s = \frac{R_L s}{R_L + s} \quad \ldots \quad (24a)$$

(The symbol $R_L||s$ denotes the value of R_L and s in parallel).

Similarly, from the input mesh:

$$v_e = i_e r + \gamma v_c = i_e r + \alpha\gamma i_e R_L^* \quad \text{(25)}$$

introducing equation (24) into equation (25) one obtains from $R_{in} = v_e/i_e$,

$$R_{in} = r + \alpha\gamma R_L^* \quad \text{(26)}$$

and defining

$$R_L^{**} = \alpha\gamma R_L^* \quad \text{(26a)}$$

$$R_{in} = r + R_L^{**} \quad \text{(26b)}$$

is obtained.

Comparison of equation (21) which is the expression of R_{in} based on the T derived equivalent circuit with equation (26) above, for the inverted π derived equivalent circuit immediately shows the superiority of the latter. The input resistance here, consists of two simple terms; the short-circuit input resistance r plus a feedback term. The fed-back term is simply the product of the effective load resistance R_L^* multiplied by the current gain and the feedback ratio. This product is the "reflexion" of R^* into the input circuit (very much like the reflexion of the secondary circuit into the primary circuit of a transformer; α and γ can loosely be compared with the effects of turns ratio and coupling coefficient.)

Equation (26) immediately provides a simple stability criterion, α, γ, R_L^* are always positive; r can be negative, and in that case, the circuit will be unstable if:

$$r + \alpha\gamma R^* < 0 \quad \text{(27)}$$

With a point contact transistor r can be negative, and the transistor circuit will be then output short-circuit unstable. This follows from the fact that for shorted output, i.e. $R_L = 0$, the input resistance will be reduced to r, i.e. become negative. Such a transistor can be output open-circuit stable provided the positive term is larger than the negative resistance r.

A similar procedure leads to the derivation of the output resistance R_{out}. Based on Fig. 3, and for generator resistance R_G in the input circuit, the output resistance can be calculated as v_c/i_c from Fig. 6. From the emitter mesh, since:

$$R_{out} = v_c/i_c \quad \text{(28)}$$

$$(R_G + r)\ i_e + \gamma v_c = 0 \quad \text{(29)}$$

therefore:

$$i_e = -\frac{\gamma c_c}{R_G + r} = -\gamma v_c/R_G^* \quad \text{(30)}$$

where:

$$R_G + r = R_G^* \quad \text{(30a)}$$

and from the collector mesh:

$$i_c = v_c/s - \alpha i_e = v_c/s + \alpha\gamma v_c/R_G^* \quad \text{(31)}$$

$$1/R_{out} = i_c/v_c = 1/s + \alpha\gamma/R_G^* \quad \text{(32)}$$

$\alpha\gamma/R_G^*$ being input conductance reflected into the output circuit

$$\quad \text{(33)}$$

Defining

$$g = 1/s = \text{open-circuit output conductance} \quad \text{(34)}$$

in this way,

$$G_G = 1/R_G \text{ and } G_G^* = 1/R_G^* \quad \text{(35), (36)}$$

$$G_{out} = g + \alpha\gamma G_G^* \quad \text{(37)}$$

The output resistance can then be represented by:

$$R_{out} = \frac{1}{1/s + \alpha\gamma/R_G^*} = s||(R_G^*/\alpha\gamma) \quad \text{(38)}$$

defining

$$R_G^{**} = R_G^*/\alpha\gamma \quad \text{(38a)}$$

one obtains:

$$R_{out} = \frac{sR_G^{**}}{s + R_G^{**}} = s||R_G^* \quad \text{(39)}$$

and it is seen that instability will occur for a negative output resistance, i.e. if:

$$s + R_G^{**} < 0 \quad \text{(40)}$$

This indicates the way in which the negative resistance component R_G^{**} is reflected into the collector mesh from the emitter mesh. Thus, the equivalent circuit and the formulæ derived, give a good picture of the interdependence of input and output mesh. In particular, it can be seen that the negative resistance r is directly responsible for the negative resistance input characteristics. The behaviour of the collector mesh is largely determined by the current amplification factor α. The constant voltage source in the emitter mesh gives an indication of the feedback voltage reflected into the emitter mesh from the output circuit.

The expression for open-circuit output condition follows directly by introducing $R_L = \infty$ into equation (26).

$$R_{\underset{\infty}{in}} = r + \alpha\gamma s \quad \text{(41)}$$

and it can be seen immediately that for the example of Fig. 4:

$$R_{\underset{\infty}{in}} = -100 + 2 \times 0{\cdot}01 \times 20\,000 = +300\Omega$$

This means that the circuit of the example is stable with the output open-circuit. It is equally effective to use the output resistance as a criterion. Thus, for a generator with $R_G = \infty$, by use of equations (30a), (38a) and (39) one obtains:

$$R_{\underset{\infty}{out}} = s(R_G^*/\alpha\gamma)$$

and since $R_G^* = \infty$ (constant current generator):

$$R_{out} = s||(R_G^*/\alpha\gamma) \quad \text{(42)}$$

For the example:

$$R_{\underset{\infty}{out}} = +20\,000\Omega$$

On the other hand, if the collector terminal is shorted to ground, the input condition is:

$$R_{\underset{0}{in}} = r \quad \text{(43)}$$

and since, in this case, r is negative, the circuit becomes unstable. Somewhere between the condition of shorted and open collector circuit, which are unstable and stable respectively, there lies a critical condition for which the input resistance becomes zero, thus:

$$R_{\underset{crit}{in}} = 0 = r + \alpha\gamma R_L^* \quad \text{(44)}$$

and for our example:

$$0 = -100 + 0{\cdot}02\,\frac{20\,000\,R_L}{20\,000 + R_L} \therefore R_L = 6\,700\Omega$$

This shows that with a load of 6 700Ω or less, the transistor circuit becomes unstable, if fed from a generator possessing no internal resistance. A small generator resistance, however, will ensure stability for the example with a 6 700Ω load. This illustrates once more the interdependence of input and output circuits. The result shows that an external load of 6 700Ω would bring the circuit to the verge of instability with a constant voltage source feeding the emitter.

Equation (42) shows that the circuit is always stable when it is fed from a constant current generator.

The voltage amplification can be calculated from:

$$A = v_c/v_G$$

where v_G = generator voltage.

Assuming unity emitter current, this gives:

$$A = \frac{\alpha R_L^*}{R_G + r + \alpha\gamma R_L^*} = \frac{1}{\dfrac{R_G^*}{\alpha R_L^*} + \gamma} \quad \ldots (45)$$

and for the example, with $R_L = 20\text{k}\Omega$,

$$R_G^* = 900, \ R_L^* = 10\,000$$

$$A = \frac{1}{900/20\,000 + 1/100} = 18$$

the voltage amplification appears as $A = 18$, which, of course, is the same as from the previous calculation.

The expressions can now be tabulated for the input and output resistances under the various conditions.

TABLE 1

Input resistance with R_L from collector to ground
$R_{in} = r + \alpha\gamma(s||R_L)$

Input resistance with collector open circuit ....
$R_{\underset{\infty}{in}} = r + \alpha\gamma s$

Input resistance with collector shorted to ground
$R_{\underset{0}{in}} = r$

Output resistance with generator possessing an internal resistance R_G ....

$R_{\underset{R_G}{out}} = s \left|\left| \dfrac{R_G + r}{\alpha\gamma} \right.\right.$

Output resistance for a perfect constant current generator, i.e., $R_G = \infty$

$R_{out} = s$

Output resistance for a perfect constant voltage generator, i.e., $R_G = 0$

$R_{\underset{0}{out}} = s \left|\left| \dfrac{r}{\alpha\gamma} \right.\right.$

Table 1 in connexion with the inverted π equivalent circuit gives a clear picture of important aspects of circuit behaviour. It is seen that the input resistance appears as a negative resistance in series with a resistance reflected from the collector mesh into the input circuit. This reflected resistance is proportional to the internal collector shunt resistance s in parallel with the load R_L. The factor $\alpha\gamma$ describes the exact amount by which this parallel combination is reduced. In the limiting cases $R_{\underset{\infty}{in}}$ and $R_{\underset{0}{in}}$ which are the input resistances for open and shorted output the reflected resistance becomes s and zero respectively. A similar mechanism applies to the output resistance. Here, the internal shunt resistance s in parallel with a reflected series combination of the negative resistance r and the generator resistance R_G, i.e. $R_G + r = R_G^*$ divided by the reflexion factor. R_G^* divided by the reflexion factor is, as it were, a resistance value reflected into the collector mesh from the emitter mesh. The limiting cases here are those where the generator internal resistances are infinity and zero respectively. In the first case the reflected resistance vanishes and one obtains s in parallel with an infinite resistance, i.e. s. In the other case, the output resistance becomes a parallel combination of r divided by the reflexion factor, with s.

Common Base Circuit With Additional External Base Resistance

Figs. 7 and 8 show a transistor circuit and its equivalent circuits with the base resistance artificially augmented by insertion of a resistance R. This circuit is of considerable practical importance. The best way of analysing the operation of this circuit is to regard it simply as a circuit containing a transistor of modified characteristics, and subsequently to ignore the implicit inclusion of the inserted base resistance R, as shown in Fig. 8(c).

For this purpose parameters r', s', α', γ' can be defined as the equivalent parameters for the "equivalent transistor". The value of these can be obtained from a combined application of the T and inverted π circuits.

Making use of the definitions given for r, s, α, γ and from equations (13) to (20):

$$\begin{aligned} r' &= r_e - (r_b + R)(\alpha - 1) \\ &= r + (\alpha - 1)\gamma s - (\gamma s + R)(\alpha - 1) \\ &= r + (\alpha - 1)(\gamma s - \gamma s - R) \end{aligned}$$

$$\therefore r' = r - R(\alpha - 1) \quad \ldots (46)$$

Thus $r' < r$, r' is more negative than r if $\alpha > 1$.

$$\gamma' = \frac{r_b + R}{r_b + R + r_c} = \frac{\gamma s + R}{\gamma s + R + (1 - \gamma)s} = \frac{1}{1 + (1-\gamma)s/(\gamma s + R)} \quad \ldots (47)$$

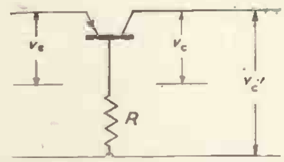
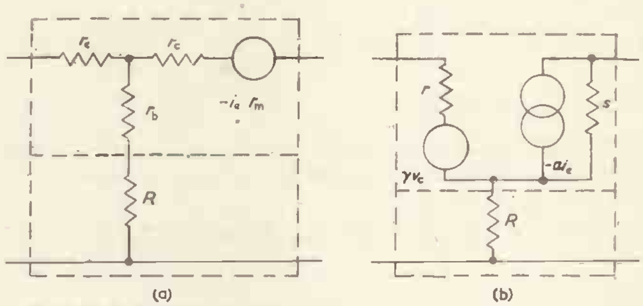


Fig. 7. Common-base circuit with additional external base resistance



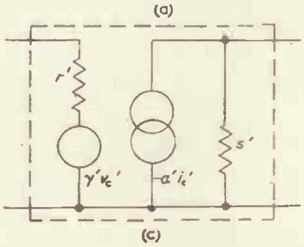


Fig. 8. Equivalent circuits for additional external base resistance

(a) T-derived
(b) Inverted π-derived
(c) "Equivalent transistor"

Thus, $\gamma' > \gamma$, i.e. the feedback is increased.

$$s' = r_b + R + r_c = \gamma s + (1 - \gamma)s + R = s + R \quad \ldots (48)$$

Thus:

$$s' > s \quad \ldots (49)$$

$$\begin{aligned} \alpha' &= (r_b + R + r_m)/s \\ &= [\gamma s + R + (\alpha - \gamma)s]/s \\ &= \gamma + R/s + \alpha - \gamma \end{aligned}$$

$$\alpha' = \alpha + R/s, \text{ i.e. } \alpha' > \alpha \quad \ldots (50)\,(51)$$

The current gain has been increased.

It is now possible to assess the effect of inserting additional base resistances of various values. To illustrate this, it is assumed that base resistances of $R = r$ and of $R = s$ are inserted respectively. Bearing in mind that these two are of widely different magnitude, a general picture can be obtained in this way.

Assuming $R = -r$:

$$r' = r + r(\alpha - 1) = \alpha r \quad \dots \quad (52)$$

$$\gamma' = \frac{\gamma s - r}{\gamma s - r + (1-\gamma)s} = \frac{\gamma s - r}{s - r} \simeq \frac{\gamma s - r}{s} \simeq \gamma - r/s \quad \dots \quad (53)$$

$$s' = s + R = s - r \simeq s \quad \dots \quad (54)$$

$$\alpha' = \alpha + R/s = \alpha - r/s \simeq \alpha \quad \dots \quad (55)$$

Assuming $R = s$:

$$r' = r - s(\alpha - 1) \simeq -s(\alpha - 1) \quad \dots \quad (56)$$

$$\gamma' = \frac{1}{1 + \dfrac{(1-\gamma)s}{\gamma s + s}} \simeq \tfrac{1}{2} \quad \dots \quad (57)$$

$$\alpha' = \alpha + 1 \quad \dots \quad (58)$$

$$s' = 2s \quad \dots \quad (59)$$

The results are summarized in Table 2. It must be borne in mind that these results are approximate, but for practical figures the errors incurred are of the order of about 1 per cent.

TABLE 2.

	R	(APPROX) $R = -r$	$R = s$
r'	$r - R(\alpha - 1)$	αr	$-s(\alpha - 1)$
γ'	$\dfrac{1}{1 + \dfrac{(1-\gamma)s}{\gamma s + R}}$	$\gamma - r/s$	$\frac{1}{2}$
s'	$s + R$	s	$2s$
α'	$\alpha + R/s$	α	$\alpha + 1$

On the basis of these approximate expressions, the input resistance and output resistance for the various terminations can now be calculated.

For $R = -r$ the following approximations result:

$$\underset{R_L}{R_{in}}(-r) = r' + \alpha'\gamma'(s'\|R_L) \simeq \alpha r + \alpha\gamma(R_L\|s) \quad \dots \quad (60)$$

$$\underset{\infty}{R_{in}}(-r) = r' + \alpha'\gamma's' \simeq \alpha(r + \gamma s) \quad \dots \quad (61)$$

$$\underset{0}{R_{in}}(-r) = r' \simeq \alpha r \quad \dots \quad (62)$$

$$\underset{R_G}{R_{out}}(-r) = s' \left\| \frac{R_G + r'}{\alpha'\gamma'} \right. \simeq s \left\| \frac{R_G + \alpha r}{\alpha\gamma} \right. \quad \dots \quad (63)$$

$$\underset{\infty}{R_{out}}(-r) = s' \simeq s \quad \dots \quad (64)$$

$$\underset{0}{R_{out}}(-r) = s'\|r'/\alpha'\gamma' \simeq s\|(r/\gamma) \quad \dots \quad (65)$$

These results are very interesting. They show that if $R = -r$ is introduced in the base lead, the input resistance (which is negative for many point contact transistors) is multiplied by α. Similarly for $R = s$:

$$\underset{R_L}{R_{in}}(s) = r' + \alpha'\gamma'(s'\|R_L) \simeq -s(\alpha - 1) + \frac{(\alpha - 1)}{2}(R_L\|2s) \quad \dots \quad (66)$$

$$\underset{\infty}{R_{in}}(s) = r' + \alpha'\gamma's' \simeq -s(\alpha - 1) + \frac{(\alpha + 1)}{2}2s = 2s \quad \dots \quad (67)$$

$$\underset{0}{R_{in}}(s) = r' \simeq -s(\alpha - 1) \quad \dots \quad (68)$$

$$\underset{G}{R_{out}}(s) = s' \left\| \frac{R_G + r'}{\alpha'\gamma'} \right. \simeq 2s \left\| \frac{R_G - s(\alpha - 1)}{\frac{1}{2}(\alpha + 1)} \right. = \frac{2}{\dfrac{1}{s} + \dfrac{\alpha + 1}{R_G - s(\alpha - 1)}} \quad \dots \quad (69)$$

$$\underset{\infty}{R_{out}}(s) = s' \simeq 2s \quad \dots \quad (70)$$

$$\underset{0}{R_{out}}(s) = s' \left\| \frac{r'}{\alpha'\gamma'} \right. = 2s \left\| \frac{-s(\alpha - 1)}{\frac{1}{2}(\alpha + 1)} \right. = 2s \frac{\dfrac{1 - \alpha}{\alpha + 1}}{\dfrac{1 - \alpha}{\alpha + 1} + 1} = -s(\alpha - 1) \quad \dots \quad (71)$$

It is interesting to observe that with a resistance of value $R = s$ introduced into the base load, the open-circuit input resistance and the open-circuit resistance are approximately equal to each other and amount to twice the internal collector shunt resistance.

These results are summarized in Table 3. It is interesting to study the figures calculated by introduction of the

TABLE 3.

	$R = O$	APPROXIMATIONS $R = -r$	$R = s$
$\underset{R_L}{R_{in}}$	$r + \alpha\gamma(R_L\|s)$	$\alpha r + \alpha\gamma(R_L\|s)$	$-s(\alpha - 1) + \dfrac{\alpha + 1}{2}(R_L\|2s)$
$\underset{\infty}{R_{in}}$	$r + \alpha\gamma s$	$\alpha(r + \gamma s)$	$2s$
$\underset{0}{R_{in}}$	r	αr	$-s(\alpha - 1)$
$\underset{R_G}{R_{out}}$	$s \left\| \dfrac{R_G + r}{\alpha\gamma} \right.$	$s \left\| \dfrac{R_G + \alpha r}{\alpha\gamma} \right.$	$\dfrac{2}{1/s + \dfrac{\alpha + 1}{R_G - s(\alpha - 1)}}$
$\underset{\infty}{R_{out}}$	s	s	$2s$
$\underset{0}{R_{out}}$	$s\|(r/\alpha\gamma)$	$s\|(r/\gamma)$	$-s(\alpha - 1)$

numerical values of the example into the expressions for inserted base resistance R of 100Ω, which is of the same order as the transistor input resistance and of 20 000Ω, which is of the same order as the collector shunt resistance. An additional value, 1 000Ω has been chosen to characterize an intermediate condition. The resulting values are tabulated in Table 3. (The simplest way of calculating these values is by determining the equivalent parameters r', s', α', γ' by use of Table 2, and to use the resulting values for introduction in the expressions of Table 1, instead of r, s, α, γ.)

Using the values of the example:

$r = -100 \qquad R_L = 20\text{k}\Omega$

$\gamma = 0{\cdot}01 \qquad R_G = 1\,000$

$\alpha = 2$

$s = 20\,000 \qquad \alpha\gamma = 0{\cdot}02 =$ "reflexion factor"

For $R = 0$ and from above and Tables 1 or 3:

$$\underset{20\text{k}\Omega}{R_{in}} = -100 + 0{\cdot}02 \times 10\,000 = 100\Omega$$

$$\underset{\infty}{R_{in}} = -100 + 0{\cdot}02 \times 20\,000 = 300\Omega$$

$$\underset{0}{R_{in}} = -100\Omega$$

$$\underset{1000}{R_{out}} = 20\text{k}\Omega\|(50 \times 900) = 20\text{k}\|45\text{k}\Omega = 13\,800\Omega$$

$$\underset{\infty}{R_{out}} = s = 20\text{k}\Omega$$

$$\underset{0}{R_{\mathrm{out}}} = 20\mathrm{k}\Omega \| -100/0{\cdot}02 = 20\mathrm{k}\Omega \| -5\mathrm{k}\Omega = -6\,700\Omega$$

For $R = 100$ Using Table 2:

$$r' \simeq -200\Omega$$
$$\gamma' \simeq 1/100 + 100/20\Omega = 0{\cdot}015 \quad \alpha'\gamma' = 0{\cdot}03$$
$$\alpha' \simeq 2$$
$$s' \simeq 20\,000\Omega$$

The values for r' s', α', γ' can now be regarded as r, s, α, γ of an "equivalent transistor." By introducing the values above into the expressions listed in Table 1, one obtains:

$$\underset{20\mathrm{k}\Omega}{R_{\mathrm{in}(100)}} = -200 + 0{\cdot}03 \times 10\,000 = 100\Omega$$

$$\underset{\infty}{R_{\mathrm{in}(100)}} = -200 + 0{\cdot}03 \times 20\,000 = 400\Omega$$

$$\underset{0}{R_{\mathrm{in}(100)}} = -200\,\Omega$$

$$\underset{1000}{R_{\mathrm{out}(100)}} = 20\,000 \| (800 \times 33) = 11\,400\Omega$$

$$\underset{\infty}{R_{\mathrm{out}(100)}} = 20\,000\Omega$$

$$\underset{0}{R_{\mathrm{out}(100)}} = 20\,000 \| (-200 \times 33) = 20\,000 \| (-6\,700)$$
$$= -10\,000\Omega$$

$R = 1\,000$ Using Table 2:

$$r' = -100 - 1\,000(2-1) = -1\,100\Omega$$

$$\gamma' \simeq 1/\left(1 + \frac{20\,000}{200 + 1\,000}\right) = 0{\cdot}057$$

$$s' = 20\mathrm{k}\Omega + 1\mathrm{k}\Omega = 21\,000\Omega$$
$$\alpha' = \alpha + 1\,000/20\,000 = 2{\cdot}05$$
$$\alpha'\gamma' = 0{\cdot}117$$

And using Table 1:

$$\underset{20\mathrm{k}\Omega}{R_{\mathrm{in}(1000)}} = -1\,100 + 0{\cdot}117(20\,000 \| 21\,000) = 100\Omega$$

$$\underset{\infty}{R_{\mathrm{in}(1000)}} = -1\,100 + 0{\cdot}117 \times 21\,000 = 1\,360\Omega$$

stable operation with output open circuit.

$$\underset{0}{R_{\mathrm{in}(1000)}} = -1\,100\Omega$$

unstable with output short circuit.

$$\underset{1000}{R_{\mathrm{out}(1000)}} = 21\mathrm{k} \| -100/0{\cdot}117 = -900\Omega$$

short circuit unstable with $1\,000\Omega$ generator.

$$\underset{\infty}{R_{\mathrm{out}(1000)}} = 21\,000\,\Omega$$

stable with constant current generator.

$$\underset{0}{R_{\mathrm{out}(1000)}} = 21\,000 \| -1\,100/0{\cdot}117 = -17\,000\Omega$$

short circuit unstable with constant voltage generator.

For $R = 20\,000\Omega$ Using Table 2:

$$r' = -20\,000\Omega$$
$$\gamma' = 0{\cdot}5$$
$$\alpha' = 3$$
$$s' = 40\,000\Omega$$
$$\alpha'\gamma' = 1{\cdot}5$$

And using Table 1:

$$\underset{20000}{R_{\mathrm{in}(20\,000)}} = -20\,000 + 3 \times 0{\cdot}5 \times (40\,000 \| 20\,000)$$
$$= -20\,000 + 20\,000 = 0$$

marginal instability

$$\underset{\infty}{R_{\mathrm{in}(20\,000)}} = -20\,000 + 3 \times 0{\cdot}5 \times 40\,000 = 40\,000\Omega$$

The circuit is stable with open output because the greatly increased feedback action exceeds the increase in negative effective emitter resistance

$$\underset{0}{R_{\mathrm{in}(20\,000)}} = -20\,000$$

The circuit is very unstable with shorted output because there is no effective feedback action to balance the large negative value of r'.

$$\underset{1000}{R_{\mathrm{out}(20\,000)}} = 40\,000 \left\| \frac{1\,000 - 20\,000}{3 \times \frac{1}{2}} \right.$$
$$= 40\,000 \| (-12\,700) = -18\,500$$

A generator with $1\,000\Omega$ internal resistance feeding the transistor would cause instability with $20\,000\Omega$ added base resistance.

$$\underset{\infty}{R_{\mathrm{out}(20\,000)}} = 40\,000\Omega$$

Fed from a constant current generator, there is a good margin of stability.

$$\underset{0}{R_{\mathrm{out}(20\,000)}} = 40\,000 \left\| \frac{-20\,000}{1{\cdot}5} \right. = -20\mathrm{k}\Omega$$

Fed from a constant voltage generator, the circuit is unstable.

These results are tabulated in Table 4.

TABLE 4.

	PARAMETER	$R=0$	$R=100$	$R=1000$	$R=20000$
	r'	-100	-200	-1100	$-20\,000$
	γ'	0.01	0.015	0.057	0.5
	α'	2	2	2.05	3
	s'	20 000	20 000	21 000	40 000
INPUT SERIES RESISTANCE	$\underset{6.7\mathrm{k}}{R_{\mathrm{in}\,(R)}}$	0	-50	-510	$-11\,400$
	$\underset{20\mathrm{k}}{R_{\mathrm{in}\,(R)}}$	100	100	100	0
	$\underset{\infty}{R_{\mathrm{in}\,(R)}}$	300	400	1 360	40 000
	$\underset{0}{R_{\mathrm{in}\,(R)}}$	-100	-200	-900	$-20\,000$
OUTPUT SHUNT RESISTANCE	$\underset{1000}{R_{\mathrm{out}\,(R)}}$	13 800	11 400	-900	$-18\,500$
	$\underset{\infty}{R_{\mathrm{out}\,(R)}}$	20 000	20 000	21 000	40 000
	$\underset{0}{R_{\mathrm{out}\,(R)}}$	$-6\,700$	$-10\,000$	$-17\,000$	$-20\,000$

Calculation of the power gain is also greatly simplified by use of the concepts established in the preceding paragraphs. It was mentioned that an analogy can be established between the transistor and a transformer; thus, the transistor and its load can be represented in the two forms

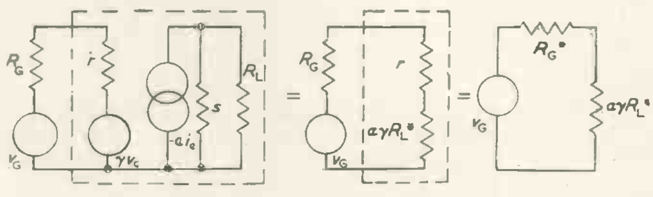


Fig. 9. Derivation of power and voltage gain

shown in Fig. 9. The chief difference here is that the effective primary and secondary impedances appear in series rather than in parallel. The series combination of $R_G + r$ being the effective generator circuit resistance R_G^* the effective emitter circuit resistance:

$$R_e = R_G^* + \alpha\gamma R_L^* \quad \ldots\ldots\ldots\ldots\ldots \quad (72)$$

can be defined as the series combination of the effective generator circuit resistance and the reflected load resistance.

The generator current i_e and power P_G can now be written down:

$$i_e = v_G/R_e \quad \ldots \quad (73)$$

$$P_G = v_G^2/R_e \quad \ldots \quad (74)$$

The available power P_L in the load follows from the collector voltage v_L which is calculated below, in terms of i_e and R_L^*:

$$v_c = -\alpha i_e R_L^* \quad \ldots \quad (75)$$

and from equation (44):

$$v_c = -\alpha R_L^* \, v_G/R_e \quad \ldots \quad (76)$$

from this:

$$A = v_c/v_G$$

$$A = -\alpha R_L^*/R_e = \text{voltage amplification} \quad \ldots \quad (77)$$

an interesting expression, proportional to the current gain and the ratio of the effective collector shunt resistance R_L^* to the effective emitter circuit resistance. (This ratio is reminiscent of the step-up ratio of a transformer.)

Thus:

$$v_c = Av_G \quad \ldots \quad (78)$$

and, breaking A down:

$$A = -\alpha R_L^*/R_e = -\alpha \frac{R_L^*}{R_G^* + \alpha\gamma R_L^*} = \frac{-\alpha}{R_G^*/R_L^* + \alpha\gamma} \quad \ldots \quad (78a)$$

And the power developed in the load is:

$$P_L = v_c^2/R_L = (-\alpha v R_L^*/R_e)^2/R_L = A^2 v_G^2/R_L \quad \ldots \quad (79)$$

The power gain is defined by:

$$G_P = P_L/P_G \quad \ldots \quad (80)$$

P_L—Power in the load

P_G—Power delivered by the source,

and from equations (74) and (79):

$$G_P = \frac{(A^2 v_G^2)R_e}{R_L v_G^2} = A^2 R_e/R_L \quad \ldots \quad (81)$$

The power amplification is therefore equal to the square of the voltage amplification times the ratio of effective emitter circuit resistance to the load resistance.

Conclusion

This analysis shows that the properties of the transistor common-base circuit are more complicated than that of a thermionic triode amplifier, even at low frequencies. This is due to the inter-dependence of emitter and collector circuits. It is, however, possible, by suitable choice of parameters and by the application of suitable equivalent circuits, to obtain a good grasp of the mechanism involved and a number of formulæ better suited for practical circuit design than those usually quoted.

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Industrial Television For British Electricity Authority

For many years past, two major problems have confronted boiler-room engineers; one was the virtual impossibility of knowing exactly what was going on inside the furnace itself; the other, how best to effect a constant supervision of the boiler water-level. Many devices have been tested to overcome these difficulties. Now, after exhaustive trials, Marconi's Wireless Telegraph Co., Ltd., have introduced an industrial television equipment which should prove a great asset to the combustion engineer, and eight such equipments are accordingly to be installed at the B.E.A.'s new power station, Willington "A" in Derbyshire.

The fuel used for heating the boilers is pulverised coal, which is jet-injected into the furnace with the air stream and is ignited by pilot oil-burners, the latter being extinguished as soon as full combustion has taken place. In such an operation, it is vital to ascertain that the fuel has fully ignited, as a firing failure, if unobserved, could result in a serious explosion.

Hitherto, observation has had to be maintained through inspection ports in the wall of the furnace, a procedure which makes supervision difficult in modern stations where control is exercised from a point which may not be immediately adjacent to the boiler. Experiments were carried out with a Marconi industrial t v. camera, fitted with a special air-and-water-cooled lens, and installed in the explosion door aperture at the base of the boiler. The television control unit and monitors were placed at a convenient point alongside the combined boiler and turbine control panel.

The experiment was successful, the B.E.A. engineers being able to follow, at the control panel, all phases of the boiler ignition procedure and to detect conditions of imperfect combustion within the area under observation.

The second experiment concerned the relatively straightforward operation of monitoring the boiler water-level gauge. For mechanical reasons, this gauge is normally located high up, near the top of the boiler, but in spite of this a constant watch has to be maintained upon it.

The camera, for the experiment, was fixed at a strategic point, televising the image of the water-gauge to the monitors at the boiler and turbine control panel.

Submerged Repeaters For The Aberdeen-Bergen North Sea Cable

The Aberdeen-Bergen submarine coaxial cable recently laid by H.M.T.S. Monarch between Scotland and Norway employs seven deep-sea repeaters laid at a maximum depth of 200 fathoms. These repeaters were built and tested in S.T.C.s new air-conditioned factory at North Woolwich which has been fully equipped with the latest test gear to enable a rigorous testing routine to be followed. Factory tested to withstand a water pressure of 5 tons/in², these torpedo-shaped repeaters have special glands at both ends for the necessary coaxial cable connexions, and can withstand the full cable tension during laying. They can, therefore, be paid out and on to the sea bed as an integral part of the cable.

The electrical design presents a number of novel features when compared with repeaters used with land coaxial cables. Like these, each is a three-stage amplifier with negative feedback, but the same amplifier is arranged to serve both directions of transmission; one direction of transmission being catered for by a group of low frequencies and the other by a higher frequency group. Power supplies for all seven repeaters on the 306-mile route are carried through the central conductor of the coaxial cable in the form of a constant direct current of 312mA fed in at the Scottish terminal at a voltage of about 1 000V.

A by-pass capacitor across the repeater circuit is arranged to give a telegraph circuit which is independent of the amplifier path.

Alarm and supervisory facilities associated with the repeaters include means of measuring the amplification up to the output of each repeater from the terminal stations. This is done by means of test frequencies for individual repeaters which are frequency-doubled after amplification and thus transmitted in the opposite direction back to the originating terminal station where they form a measure of the degree of amplification effected.

Terminal equipment includes frequency translating equipment for 36 telephones circuits, i.e. three groups of twelve circuits each occupying the frequency range 60 to 108kc/s.

Gain Measurements on Computing Amplifiers

By A. B. Johnson*, B.Sc., A.M.Brit.I.R.E.

Computing amplifiers used in d.c. analogue computors commonly have very high gain factors, tens of thousands to tens of millions, in order to achieve the " virtual earth " necessary for accurate mathematical operations. Because the pass-band extends to zero frequency, special techniques are often necessary in the measurement of gain.

THE internal gain of a computing amplifier determines the computing accuracy, output impedance and grid terminal impedance. It also affects the decay time-constant of a " holding " integrator. It is a simple matter to calculate the gain, but since large errors are likely, the results should be checked by measurement.

To measure this gain is not easy since very often the amplifier cannot be operated without negative feedback. An amplifier with an internal gain of $\times 10^5$ and a short-term drift of 1mV (referred to input grid) might produce an output noise level of 100V which would probably overload the output stage before any test signal could be applied.

Methods of measurement can be divided roughly into two classes:

(a) DIRECT (1, 2, 3, below)

Means are found to reduce the drift output without affecting the test signal. For the latter, the amplifier is effectively working without feedback. Since the drift is of zero or very low frequency, the test signal is preferably a sinusoid of suitably higher frequency.

(b) INDIRECT (4, 5, 6, 7, below)

The drift output is small because of negative feedback and the gain is deduced from some property of the experimental arrangement.

The best method to use in any particular case depends upon the gain and drift and if these are both low enough to permit the amplifier to be used without feedback, measurement is quite simple. In general, the higher the gain, the more difficult it is to make an accurate measurement. The difficulties introduced by drift depend upon the method used; in some cases it is a limiting factor, in others it is unimportant.

Direct Methods

(1) INTRODUCTION OF A.C. COUPLINGS

The drift noise output on open loop can be reduced to a tolerable level if one or more direct couplings in the amplifier are temporarily replaced by a.c. couplings. For instance, the typical direct coupling in Fig. 1(a) can be altered to Fig. 1(b).

The time-constant *CR* is chosen to be sufficiently large to pass the test signal without attenuation, while *R* is made large enough to produce negligible loading on the previous circuit. The bias voltage is adjusted to preserve the designed operating conditions for the second stage. This method is inconvenient and cannot be used to take frequency response curves since the stray capacitances are considerably modified.

(2) FEEDBACK AT D.C. ONLY

Consider the simple differentiator in Fig. 2

The resistor *R* provides 100 per cent feedback at d.c. The

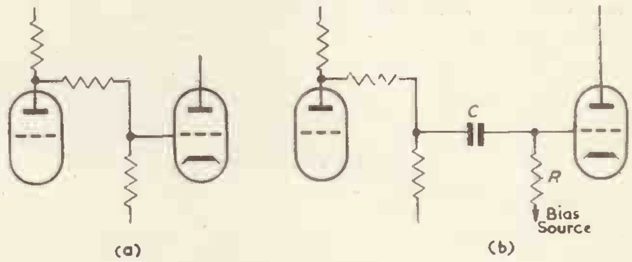

Fig. 1. Introduction of a.c. couplings

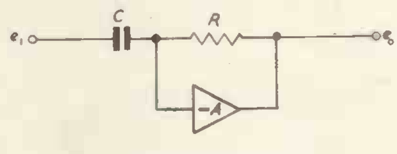

Fig. 2. Feedback at d.c.

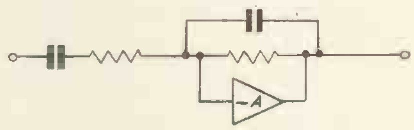

Fig. 3. Prevention of excessive phase shift at high frequencies

transfer function is given by:

$$e_o/e_i = \frac{-AR}{R + (A+1)\,1/pC}$$

or for sinusoids:

$$e_o/e_i = \frac{-AR}{R - j\,(A+1)/\omega C}$$

and:

$$|e_o/e_i| = \frac{-AR}{[R^2 + (A+1)^2/\omega^2C^2]^{\frac{1}{2}}}$$

Now if $R \gg (A+1)/\omega C$, $|e_o/e_i| = -A$.

Hence, if for the test frequency, ω, the capacitive reactance is much smaller than $R/(A+1)$, the amplifier behaves

* *Saunders Roe Ltd.*

as if there were no feedback at this frequency. For example, let $A = 10^5$, $R = 10^7\Omega$, $\omega = 2\pi \cdot 10^3$, then $C \gg 1{\cdot}6\mu F$. Quite accurate results will be obtained for $C = 16$ to $32\mu F$, but note that the connexion shown in Fig. 2 may be inadmissible because of system instability. This depends upon the amplifier phase and gain characteristics. The circuit might have to be altered to Fig. 3, where the additional components prevent excessive phase shift at high frequencies. The results would in this case require slight modification.

A more refined method is shown in Fig. 4.

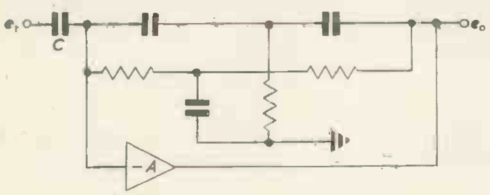

Fig. 4. Development of Fig. 3 using twin-*T* network

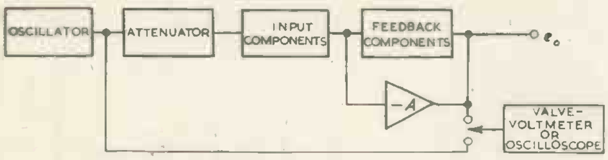

Fig. 5. Experimental set-up

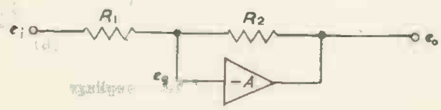

Fig. 6. Grid voltage method

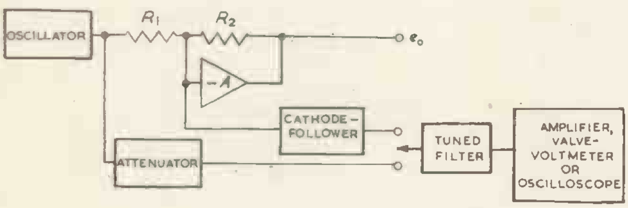

Fig. 7. Experimental layout for Fig. 6

The feedback network is still resistive and gives 100 per cent feedback at d.c., but the twin-*T* configuration ensures that there is no test signal current fed back to the grid node. The value of C then can be considerably reduced. The experimental set-up would be as in Fig. 5.

(3) Grid Voltage

Consider the arrangement in Fig. 6.

We have:

$$e_o/e_i = -R_2/R_1$$

(error < 1 per cent for $A > 200$, $R_2/R_1 = 1$)

and:

$$e_g = -e_o/A = e_i \cdot R_2/A \cdot R_1$$

Comparison of e_g with e_i will give A.

The experimental layout is shown in Fig. 7.

This method is convenient and more suitable for taking frequency response curves than methods (1) and (2).

(4) Computation Error

Referring to Fig. 6:

$$e_o'/e_i \quad \frac{AR_2}{R_2 + (A+1)R_1} = m \simeq R_2/R_1 = k = e_o/e_i$$

If $e_o'/e_i = m$ is measured and $R_2/R_1 = k$ is known:

$$A = \frac{m\,(k+1)}{k-m}$$

The difficulty here is that $k - m$ is very small compared with m or k if A is large and it may not be possible to measure it sufficiently accurately. For instance, if $A = 10^5$, $k = 10^3$, then $k - m = 10$. For 10 per cent accuracy in the estimate of A, k and m must be known to within 0·05 per cent and therefore k should be made as large as the drift will permit. This method is suitable only for values of A up to about 2 000.

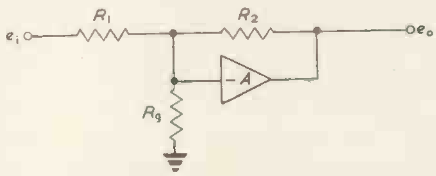

Fig. 8. Grid terminal impedance

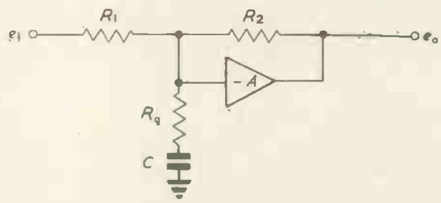

Fig. 9. Modification of Fig. 8

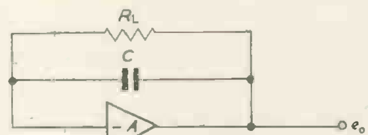

Fig. 10. Integrator decay time-constant

(5) Grid Terminal Impedance

The circuit in Fig. 8 has a transfer function:

$$e_o/e_i = \frac{AR_2}{R_1(1+A) + R_2 + (R_1R_2/R_3)}$$

$$\simeq \frac{-AR_2}{(A+1)R_1 + R_1R_2/R_g}$$

Now letting $R_g = R_2/A+1$ gives:

$$e_o/e_i = \frac{-AR_2}{2\,(A+1)\,R_1} \simeq -R_2/2R_1$$

Thus by shunting the grid with a resistance equal to $R_2/A+1$ the output is reduced from $-e_i \cdot R_2/R_1$ to half this value. If R_g is adjusted until the output voltage falls to half and R_g then measured we have:

$$A = \frac{R_2}{R_g} + 1 \simeq \frac{R_2}{R_g}$$

Unfortunately the drift output in the case of the unshunted grid is $(1 + R_2/R_1)\delta$ (where δ is the drift referred to the input grid) and with the grid shunted the drift becomes $-A/2 \, . \, \delta$. Thus the gain and drift jointly impose limitations on this method. However, by a simple modification this difficulty can be overcome. In Fig. 9, C has been added to eliminate the grid shunting effect at zero and very low frequencies, while its reactance is negligible compared with R_g at the test frequency.

(6) Integrator Decay Time-constant

If the capacitor of an integrating circuit (Fig. 10) is charged to a voltage E and the grid circuit is disconnected, the output voltage is a function of time:

$$e_o = E \, . \, e^{-t/T_1}$$

where T_1 is the leakage time-constant CR_L.

Turning now to Fig. 11, where R_L is assumed infinite and R_g has been added:

$$e_o = -E \, . \, e^{-t/T_2}$$

where $T_2 = ACR_g$.

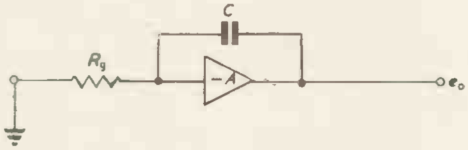

Fig. 11. Modification of Fig. 10

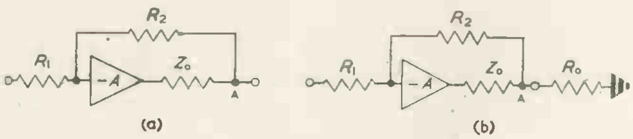

Fig. 12. Output impedance

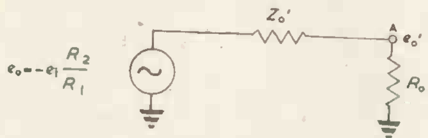

Fig. 13. Equivalent circuit of Fig. 12(b)

In the second case the decay is dependent upon A and if the time-constant can be evaluated experimentally A can be calculated from $A = T_2/CR_g$.

Now in order to isolate the two decay effects due to R_L, R_g, the first should be estimated by finding T_1, in the absence of R_g. Then a value of R_g should be chosen that $AR_g \ll R_L$ and the effect due to R_L can then be ignored.

For a small change Δ in e_o, occurring in time t, $\Delta \simeq Et/T_2$. Hence $A \simeq e_o \, . \, t/CR_g \, . \, \Delta$.

This method is convenient and valid provided that the drift output, $A\delta(1 - e^{-t/T_2})$ does not introduce serious error. In order that this should not be so we require $A\delta \ll E$. So we see that this method is limited to amplifiers whose gain-drift products are a small fraction of their maximum output voltage excursions, in which case, of course, the gain can be measured directly in open loop. The advantage of this method is its simplicity.

(7) Output Impedance

This parameter of the circuit shown in Fig. 12(a) is given by:

$$Z_o' = \frac{(R_1 + R_2)Z_o}{R_2 + (A + 1)R_1} \simeq \frac{(R_1 + R_2)Z_o}{AR_1}$$

where Z_o is the output impedance of the amplifier output stage.

If, as in Fig. 12(b) we shunt the output with a load R_o the equivalent circuit becomes as in Fig. 13.

Now:

$$e_o' = -e_1 \, . \, R_2/R_1 \, . \, \frac{R_o}{Z_o + R_o} = \frac{e_o \, . \, R_o}{\dfrac{Z_o(R_1 + R_2)}{R_2 + (A + 1)R_1} + R_o}$$

Whence:

$$A = Z_o/R_o \, . \, \frac{R_2 + R_1}{R_1} \, . \, \frac{e_o'}{e_o - e_o'} - R_2/R_1 - 1$$

Z_o can be found with sufficient accuracy by calculation, though the results will be more accurate if Z_o is increased artificially with an external resistance of known value. The value of R_o must not be too low as to cause overloading of the output stage and in fact $R_o \leqslant Z_o \, . \, e_o/E - e_o$, where e_o is

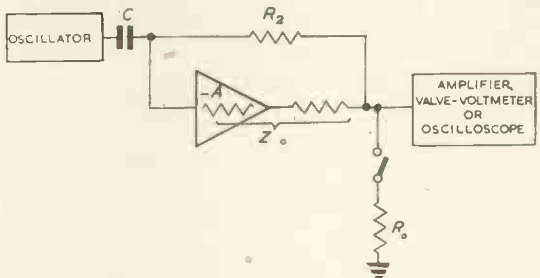

Fig. 14. Development of output impedance method

the output voltage (at point A) in the absence of load R_o and E is the amplifier terminal maximum excursion. We have:

$$e_o - e_o' = e_o \, . \left[1 - \frac{R_o}{\dfrac{Z_o(R_1 + R_2)}{R_2 + (A + 1)R_1} + R_o} \right]$$

and:

$$e_o = \frac{E \, . \, R_o}{Z_o + R_o}$$

$$\therefore \; e_o - e_o' = \frac{E \, . \, n \, . \, k}{(n + 1)(kn + 1)}$$

where $n = Z_o/R_o$, $k = \dfrac{R_1 + R_2}{R_2 + (A + 1)R_1}$

For a given E, $e_o - e_o'$ is a maximum when $k = n = 1$.

Then $e - e_o' = E$, $R_o = Z_o$, $\dfrac{R_1 + R_2}{R_2 + (A + 1)R_1} = 1$.

The last equation implies that $R_2 \gg (A + 1)R_1$ and the drift output will approach the value $A\delta$. Thus for the method to succeed we must have $A\delta \ll E/4$.

This restriction may be removed by using an a.c. signal as in Fig. 14. The effective value of k is nearly unity for the signal and very small for the drift. The reactance of C to the test signal frequency is $\ll R_1(A + 1)$. As in Fig. 4, R_2 may, with advantage, be replaced by a rejector network tuned to the test frequency.

An Approximate Treatment of Cascaded Four-Terminal Networks

By H. L. Armstrong*

An approximate expression is derived for the n^{th} power of a 2×2 matrix. The result is used in an approximate treatment of a ladder network used as a filter.

WHEN three (or four) terminal networks are represented by their transmission matrices θ, which express the relation between voltages and currents at the two ends of the network through:

$$\begin{bmatrix} V_1 \\ i_1 \end{bmatrix} = \theta \begin{bmatrix} V_2 \\ i_2 \end{bmatrix} = \begin{bmatrix} \theta_{11} & \theta_{12} \\ \theta_{21} & \theta_{22} \end{bmatrix} \begin{bmatrix} V_2 \\ i_2 \end{bmatrix} \quad \ldots\ldots (1)$$

the network formed by cascading such sections has a matrix which is the product of the matrices of the individual sections. In particular, if n identical sections are cascaded, the matrix of the resulting network is the n^{th} power of the matrix of one section[1,2].

Even with the condensed expressions which have been

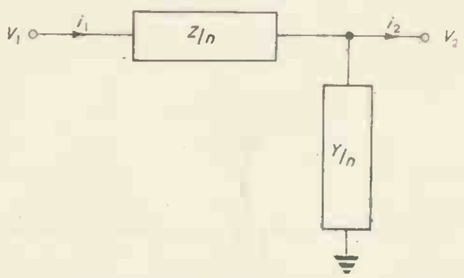

Fig. 1. Diagram of the *L*-section used as an example

given[3,4], the power of the matrix is still rather complicated. It is the present purpose to develop approximate expressions for the n^{th} power of such a matrix. The illustrations will deal with a ladder network, i.e., a cascade of L-sections, but the methods developed are equally applicable to other types of section. The L-section has shunt admittance Y/n, and series impedance Z/n, so that Y is the total admittance in the network, and Z the total impedance.

The treatment is based on the two relations:

$$(1 + x/n)^n \sim e^x\, e^{-x^2/2n} \quad (|x/n| < 1) \quad \ldots\ldots\ldots (2)$$

and

$$e^A = \frac{1}{\lambda_1 - \lambda_2}\left\{ e^{\lambda_2}(\lambda_1 I - A) - e^{\lambda_1}(\lambda_2 I - A)\right\} \ldots\ldots (3)$$

A being a 2×2 matrix, I the unit matrix of order 2, and λ_1 and λ_2 the Eigen values of A ($\lambda_1 \neq \lambda_2$). Note that (2) is true for matrices, as well as for numbers. These relations are established in Appendices 1 and 2, respectively.

It is convenient to use the alteration matrix A, defined by:

$$1/n\,(A) = \begin{bmatrix} a_{11}/n & a_{12}/n \\ a_{21}/n & a_{22}/n \end{bmatrix} = \theta - I = \begin{bmatrix} \theta_{11} - 1 & \theta_{12} \\ \theta_{21} & \theta_{22} - 1 \end{bmatrix} \ldots (4)$$

since from equation (2):

$$\theta^n = (I + A/n)^n \sim e^A\, e^{-A^2/2n} \quad \ldots\ldots\ldots (5)$$

For the L-section considered:

$$A = \begin{bmatrix} YZ/n & Z \\ Y & 0 \end{bmatrix} \ldots\ldots\ldots\ldots (6)$$

For n large (the network divided into many sections) the term YZ/n will be neglected. Then:

$$A \approx \begin{bmatrix} 0 & Z \\ Y & 0 \end{bmatrix} \text{ also } \lambda_1 = \sqrt{(YZ)},\ \lambda_2 = -\sqrt{(YZ)}$$

and equation (3) gives:

$$\theta^n \approx \exp \begin{bmatrix} 0 & Z \\ Y & 0 \end{bmatrix} = \begin{bmatrix} \cosh \sqrt{(YZ)} & \sqrt{(Z/Y)} \sinh \sqrt{(YZ)} \\ \sqrt{(Y/Z)} \sinh \sqrt{(YZ)} & \cosh \sqrt{(YZ)} \end{bmatrix} \ldots\ldots\ldots (7)$$

But the matrix obtained is just that of a transmission line, to which the network reduces as the number of sections into which it is divided becomes larger and larger. It has been shown before[5] how a ladder or similar network reduces to a transmission line as the number of subdivisions becomes infinite, but the above treatment is believed to have some novelty. It is also interesting to notice that, if Y and Z are the admittance and impedance per unit length, the transmission line equations can be written[6]:

$$\frac{d}{dx}\begin{bmatrix} V \\ i \end{bmatrix} + \begin{bmatrix} 0 & Z \\ Y & 0 \end{bmatrix} \begin{bmatrix} V \\ i \end{bmatrix} = 0 \quad \ldots (8)$$

i.e., the transmission line behaviour is merely the law of natural growth, but in terms of matrices rather than numbers.

The present purpose, however, is to carry the approximation beyond that obtained in equation (7). From equation (6):

$$A - A^2/2n = \begin{bmatrix} YZ/2n - Y^2Z^2/2n^3 & Z - YZ^2/2n^2 \\ Y - Y^2Z/2n^2 & -YZ/2n \end{bmatrix}$$

$$\approx \begin{bmatrix} YZ/2n & Z - YZ^2/2n^2 \\ Y - Y^2Z/2n^2 & -YZ/2n \end{bmatrix} \ldots (9)$$

If powers of $1/n$ higher than the second are neglected. The Eigen values of the approximate expression for $A - A^2/2n$ are:

$$\lambda_1 = \sqrt{(YZ - 3Y^2Z^2/4n^2)},\ \lambda_2 = -\sqrt{(YZ - 3Y^2Z^2/4n^2)} \quad \ldots\ldots\ldots (10)$$

Again powers of $1/n$ higher than the second are neglected. When these values and the approximation to $A - A^2/2n$ are put in equation (3) the result is:

$$\theta^n \approx \exp(A - A^2/2n) \approx I \cosh \sqrt{(YZ - 3Y^2Z^2/4n^2)} + \frac{A - A^2/2n}{\sqrt{(YZ - 3Y^2Z^2/4n^2)}} \sinh \sqrt{(YZ - 3Y^2Z^2/4n^2)}$$

* *Pacific Semiconductors Inc. U.S.A.*

$$= \begin{bmatrix} \cosh\Omega + 1/2n\sqrt{\left(\frac{YZ}{1-3YZ/4n^2}\right)}\sinh\Omega & \sqrt{\left(\frac{Z}{Y}\right)}\frac{1-YZ/2n^2}{\sqrt{(1-3YZ/4n^2)}}\sinh\Omega \\ \sqrt{\left(\frac{Y}{Z}\right)}\frac{1-YZ/2n^2}{\sqrt{(1-3YZ/4n^2)}}\sinh\Omega & \cosh\Omega - 1/2n\sqrt{\left(\frac{YZ}{1-3YZ/4n^2}\right)}\sinh\Omega \end{bmatrix} \ldots (11)$$

with the notation $\Omega = \sqrt{(YZ - 3Y^2Z^2/4n^2)}$

As $n \to \infty$ equation (11) reduces to equation (7), as it should.

As an example, let $Y = j\omega C$ and $Z = j\omega L$. Let the network be connected between a generator of internal impedance $R_o = \sqrt{(L/C)}$ and internal voltage V_o, and a load resistance $R_L = R_o$, across which a voltage V_L is developed. Then, in terms of elements of θ^n:

$$V_L/V_o = \frac{1}{(\theta^n)_{11} + R_o(\theta^n)_{21} + (\theta^n)_{12}/R_o + (\theta^n)_{22}} \ldots (12)$$

in this case this gives:

$$V_L/V_o = \frac{1}{2\cosh\Omega + 2\dfrac{1 - YZ/2n^2}{\sqrt{(1-3YZ/4n^2)}}\sinh\Omega} \ldots (13)$$

When absolute values are taken:

$$|V_L/V_o| = 1/2\sqrt{\left[1 + (\omega^2LC/4n^2)\left(\frac{1+\omega^2LC/n^2}{1+3\omega^2LC/4n^2}\right)\sin^2(\omega\sqrt{(LC)}\sqrt{(1+3\omega^2LC/4n^2)}\right]} \ldots (14)$$

Fig. 2 shows this ratio, plotted against ω/ω_o, ($\omega_o = 2n/\sqrt{(LC)}$), for $n = 10$, i.e., for a network with total

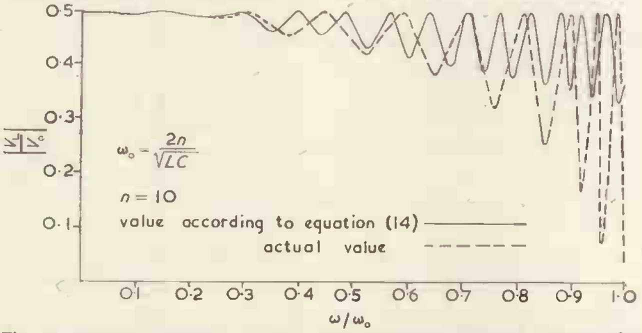

Fig. 2. The ratio $|V_L/V_o|$ compared with the actual value of $|V_L/V_o|$ for a network of 10 sections plotted against the "normalized frequency" ω/ω_o

inductance L and capacitance C, divided into 10 sections. The approximation is seen to be good for values of ω up to about $\omega_o/2$.

Conclusions

An approximate expression for the n^{th} power of 2×2 matrix has been derived. This expression has been used in an approximation to the actual behaviour of a filter network. Other circuits could be treated similarly.

Appendix 1

Since $\ln(1 + x/n) = x/n - x^2/2n^2 + \ldots$

$$(1 + x/n)^n = \exp[n\ln(1 + x/n)] \approx$$
$$\approx \exp(x - x^2/2n) = e^x\exp(-x^2/2n) \ldots (15)$$

This approximation is true for $|x/n| < 1$, and for values a little less than one, the agreement is quite good. For any x, the approximation becomes better and better as n increases, becoming perfect as $n \to \infty$. In this sense, the approximation is better than that obtained by using the first two or three terms of a binomial expansion. In this case, the error approaches a value independent of n, as $n \to \infty$.

Appendix 2

Consider a polynomial $P_m(A)$ of a 2×2 matrix A, with Eigen values λ_1 and λ_2, $(\lambda_1 \neq \lambda_2)$, given by:

$$P_m(A) = \sum_{k=0}^{m} 1/k!\ A^k \ldots (15)$$

By Sylvester's theorem[7]:

$$P_m(A) = \frac{1}{\lambda_1 - \lambda_2}\left\{P_m(\lambda_2)(\lambda_1 I - A) - P_m(\lambda_1)(\lambda_2 I - A)\right\} \ (17)$$

$P_m(\lambda)$ indicates the corresponding scalar polynomial.

Also, by definition:

$$e^A = \sum_{k=0}^{\infty} 1/k!\ A^k \ldots (18)$$

Now:

$$e^A - P_m(A) = \sum_{k=m+1}^{\infty} 1/k!\ A^k \ldots (19)$$

Let a_u be an upper bound to the elements of A. Then $2^{n-1}a_u^n$ is an upper bound to the elements of A^n, as is obviously $(2a_u)^n$. Also since $(m + k)! > (m!)(k!)$, every element in the matrices in the summation in equation (19) is dominated by the corresponding element of the series:

$$\frac{(2a_u)^{m+1}}{(m+1)!}\sum_{k=0}^{\infty}\frac{(2a_u)^k}{k!}\begin{bmatrix}1 & 1\\ 1 & 1\end{bmatrix} \ldots (20)$$

But each element in the matrix represented by this sum is just $e^{2a_u}(2a_u)^{m+1}/(m + 1)!$ which approaches zero as m approaches infinity.

Accordingly:

$$\lim_{m \to \infty} P_m(A) = e^A \ldots (21)$$

and this in equation (16) gives:

$$e^A = \frac{1}{\lambda_1 - \lambda_2}\left\{e^{\lambda_2}(\lambda_1 I - A) - e^{\lambda_1}(\lambda_2 I - A)\right\} \ (22)$$

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Pegasus — A New Computor

Ferranti Ltd., have recently announced "Pegasus," the Ferranti Packaged Computor No. 1. This is a medium-sized general-purpose machine suitable for carrying out the type of calculations which arise in industry, research and management. The complete computor is built up from a range of standardized plug-in units.

Input is by means of punched paper tape at a maximum speed of 200 characters/second and the output punched tape at 15 characters/second or by automatic typewriter.

Nickel delay lines are used for the computing store and a magnetic drum for the main store.

Notes from North America

IRE News

At the January meeting of the Board of Directors of the Institute of Radio Engineers, the following were appointed to the Board for 1955:

W. R. G. Baker—Treasurer (Vice-President of Electronics, General Electric Company)
H. Pratt—Secretary
J. R. Pierce—Editor (Director of Electronics Research, Bell Telephone Laboratories)
A. N. Goldsmith—Editor Emeritus of the IRE
A. V. Loughren (Director of Research, Hazeltine Corporation)
H. Vollum (President, Tektronix, Inc.)

The 1955 IRE National Convention is to be held from 21-24 March in New York City. There will be 55 technical sessions, including two special symposia on "Magnetic Recording for the Engineer" and "Trends in Automatization of Procedures and Processes in Business and Industry". There will also be a Radio Engineering Show containing some 700 exhibits.

Versatile Induction Heater

The Westinghouse Electric Corporation of Baltimore has recently supplied a versatile induction heating equipment to an American manufacturer of agricultural machinery for the hardening shafts and gears.

The equipment, shown in the illustration, consists of two 50kW, 450kc/s generators with work tables, water cooling system and auxiliaries.

Each of the work tables holds two rotating lift spindles which are manually loaded and used for the hardening of small shafts and other cylindrical components. If required, the components can be selectively hardened to enable subsequent drilling operations to be carried out in adjacent unhardened areas. Under these conditions the output varies from 120 to 300 components an hour depending on the size of the component.

The generators can be paralleled for the heat treatment of the larger items such as gears.

For this operation, the centre station shown in the illustration is used and the cycle is as follows:

1. Preheat treatment up to 50kW for about 15 seconds.

2. Air cooling or soaking for about 10 seconds.
3. High power heat treatment at full output of 100kW for 7-10 seconds for contour or surface hardening.

With this sequence of operation the output is 120 gears an hour.

New Valves

Three new valves, types 5U4-GB, 6AF4-A, 6CG7, have been announced by the Tube Division of the Radio Corporation of America, Harrison, New Jersey.

The 5U4-GB is a full-wave vacuum rectifier of the glass-octal type intended for use in the power supplies of television receivers and in radio equipment having high d.c. requirements.

In comparison with the 5U4-G, the new 5U4-GB has the same maximum voltage ratings but higher current ratings. The 5U4-GB is rated to withstand a peak anode current of 1·0A per anode, a value 48 per cent higher than for the 5U4-G. For the same applied alternating voltage, the 5U4-GB can deliver a d.c. output current approximately 22 per cent higher with capacitor-input circuits, and 28 per cent higher with choke-input circuits.

The 6AF4-A is a 7-pin miniature triode designed especially for use as an oscillator in tuners of u.h.f. television receivers covering the range from 470 to 890Mc/s. It is similar to the 6AF4, but is $\frac{3}{8}$in shorter to permit more compact tuner designs.

Like the 6AF4, the 6AF4-A has good frequency stability and features a small mount structure with small elements to provide low interelectrode capacitances; short internal leads to reduce lead inductance and resistance; silver-plated base pins to minimize losses caused by skin effect at the ultra-high frequencies; and double base-pin connexions for both anode and grid. The double connexions are arranged so as to facilitate use of the 6AF4-A with either series or parallel resonant lines and to offer greater flexibility in circuit connexions.

The 6CG7 is a 9-pin miniature version of the 6SN7-GT and is intended for use particularly as a vertical deflexion oscillator and horizontal deflexion oscillator in television receivers. This type is designed with a 600mA heater having a controlled warm-up time to ensure dependable performance in television receivers employing a single series-connected heater chain.

Design features of the 6CG7 include a structure which permits cool operation of the grids with the result that emission from them is minimized. The structure also incorporates an internal shield which provides effective shielding between the triode units to prevent electrical coupling between them.

The 6CG7 may also be used as a phase invertor, multivibrator, synchronizing separator and amplifier, and resistance-coupled amplifier.

Webster Chair of Electrical Engineering

Dr. Robert A. Ramey, Jr., Manager of the Magnetic Development Section of the Materials Engineering Department of the Westinghouse Electric Corporation (Pittsburgh), has been appointed Visiting Webster Professor of Electrical Engineering at the Massachusetts Institute of Technology for the second semester of the current academic year. Dr. Ramey will participate in the development of teaching and research in the new area of solid state non-linear devices and their applications to power modulators.

Short News Items

The Radio Industry Council announces the award of premiums of 25 guineas each for articles published in the public technical press during 1954 as follows:

Two Premiums for the three articles:

"LEO Lyons' Electronic Office," by J. M. M. Pinkerton and E. J. Kaye (*Electronic Engineering*, July, 1954).

"LEO Operation and Maintenance," by E. H. Lenaerts (*Electronic Engineering*, August, 1954).

"LEO A Checking Device for Punched Data Tapes," by E. J. Kaye and G. R. Gibbs (*Electronic Engineering*, September, 1954).

One Premium for author of the two articles:

"Wide-Band I. F. Amplifiers," by H. S. Jewitt (*Wireless World*, February, 1954).

"Feedback I. F. Amplifier for Television," by H. S. Jewitt (*Wireless World*, December, 1954).

One Premium for author of the two articles:

"High-Speed Magnetic Amplifiers," by A. E. Maine (*Electronic Engineering*, May, 1954).

"Three-Phase High-Speed Magnetic Amplifiers," by A. E. Maine (*Electronic Engineering*, December, 1954).

One Premium for:

"A Torquemeter for Testing Gas Turbine Components," by J. F. Field and D. H. Towns (*Electronic Engineering*, November and December, 1954).

One Premium for:

"Dip-Soldered Chassis Production," by W. R. Cass and R. M. Hadfield (*Wireless World*, November, 1954).

The presentation of the awards will take place at a luncheon to be held by the Public Relations Committee of the Radio Industry Council on Thursday, 10 March.

The Mullard Company have opened a new Electronics Display Centre at their Publicity Offices in Gerrard Place, London, W.1. In this Centre, the Mullard Products are displayed and explained in a comprehensive way. An attempt has been made to provide a setting in harmony with electronic products and their fields of application. The showroom is on two floors. On the first, examples from each group of the company's products are arranged in separately illuminable panels, and coloured wall maps show the ramifications of the Mullard Company at home and overseas. Current Mullard literature is also on display. On the lower floor twenty-seven panels are used to display more Mullard products. Also on this floor are housed larger exhibits which are featured from time to time in connexion with a planned programme of exhibitions and live demonstrations. In the same building there is a cinema and conference room. Visits to this showroom will be encouraged by the Mullard Company. These will extend to parties from education establishments, students, apprentices and trainees, business contacts of the various company divisions, Government Departments, the radio industry and trade, etc.

The General Electric Co., Ltd., have introduced a new method of training boys leaving public and grammar schools to become professional engineers capable of filling responsible engineering, administrative and sales positions in industry. It is operated by the Company in conjunction with the Birmingham College of Technology. In future, all boys in this category who are selected by the Company for training will, on leaving school at the age of eighteen, be given a five-year course, in which full time sessions of six months at the College of Technology will alternate with six month periods of industrial training in the Company's works. During the whole of the five-year course, the Company pays a student's college fees and also gives him suitable remuneration to cover living expenses. Two country houses in the Midlands, Castle Bromwich Hall, near Birmingham, and Coombe Abbey, near Coventry, are used by the Company as Halls of Residence.

Marconi's to Supply Radio Beacons. In compliance with regulations laid down at the 1951 Conference for the Reorganization of Maritime Radio Beacons, the Portuguese Lighthouse Department are to duplicate their existing radio beacons. To this end they have now placed an order with Marconi's Wireless Telegraph Co. Ltd., for four duplicate equipments of the latest design. Radio beacons are in many respects equivalent to lighthouses, but instead of a beam of light they transmit wireless signals from which vessels with direction-finding equipment can obtain their bearings. Their advantage over coastal warning lights is that they are unaffected by fog.

Baird & Tatlock Ltd. have recently opened new offices and showrooms in Manchester covering a complete range of scientific instruments, apparatus, laboratory fittings and chemicals. The Manchester office will be able to deal directly with orders from laboratories in the North of England and, in general, provide many of the facilities previously only available in London. The manager of the new branch will be Mr. A. E. Hobson and the address is 58 Lever Street, Manchester, 1.

The firm Benigno Novoa, of Bolivar 1061, Buenos Aires, Argentina, offer their services as agents in the Argentine Republic to British manufacturers and exporters.

Ferranti Ltd., Edinburgh, who have been engaged in the development of Resin Cast Transformers and Chokes for the past four years, have now received a Limited Type Approval Certificate from the Radio Components Standardization Committee for these components. The new range of transformers and chokes have been named the Pentland Series.

The Council of The Institution of Electrical Engineers have elected His Royal Highness the Duke of Edinburgh to Honorary Membership.

Electrical & Musical Industries Ltd. announce the change of name of Emitron Television Ltd. to E.M.I. Electronics Ltd. The new company will be a controlling and co-ordinating company, and will absorb as subsidiaries the three existing companies, E.M.I. Engineering Development Ltd., E.M.I. Factories Ltd., and E.M.I. Research Laboratories Ltd. E.M.I. Electronics Ltd. will be concerned with the design, development and marketing of all electronic devices other than those for Government requirements, utilizing for this purpose the extensive resources of its three subsidiary companies.

The United Kingdom Atomic Energy Authority announce the appointment of Colonel G. W. Raby, C.B.E., as Deputy Director (Engineering) to the Research Group at Harwell. Mr. D. E. H. Peirson has been appointed as Secretary to the Authority.

Chapman & Hall, Ltd., have been appointed sole British agents for the book "Fundamentals of Transistors," by L. M. Krugman, published in the United States of America by the John F. Rider Co., Inc.

The Saar Trade Fair will be held in Saarbrucken from 23 April to 8 May. Further information may be obtained from the Fair Management, Saarbrucken, Messegelande.

Metropolitan-Vickers Electrical Co. Ltd. have appointed Mr. R. H. Kelsall as Assistant Chief Electrical Engineer, Electronic Control Department. This follows the transfer of Mr. S. A. Ghalib to be Group Leader of the A.E.I.-John Thompson Industrial Atomic Energy Group.

Mr. W. A. Flint has resigned from Sunvic Controls Ltd., and is taking up a position as Publicity Officer to Petters Ltd., Staines.

LETTERS TO THE EDITOR

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

Amplifier Output Impedance

DEAR SIR,—The letter in your December issue from Messrs. McDonnell and Earnshaw brings, very forcibly, to my attention, the dangers of accepting approximate and empirical formulae in connexion with electronic circuits. It is of course easy to use a text book answer, but in many cases, of which the amplifier with or without feedback is a notable one, an exact statement of conditions is often more easily grasped and more easy to apply.

Firstly, an amplifier can be shown as a T or π network of impedances, and therefore may be considered as a three terminal network which, if we assume a value for any one impedance of the three, may be a passive network, the only applied potential being the input potential to the grid. This point is mentioned in passing, to demonstrate that ordinary network theorems may profitably be applied to electronic circuits.

Now, assume an amplifier to be a four terminal network such that $e_{3\text{-}4} = me_{1\text{-}2}$, and that 2 and 4 are connected (Fig. 1).

If we consider 1 as the input grid of an amplifier, and 4 as the output terminal, then, knowing m, and any impedance of the T or π, the amplifier is exactly calculable.

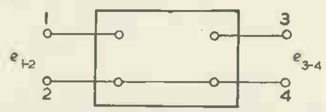


Fig. 1. Amplifier as four terminal network

Now, consider an amplifier with an external T network connected as shown in Fig. 2.

Since the amplifier itself is in no way altered, then $e_{3\text{-}4} = me_{1\text{-}2}$ as before, and we may conveniently let $e_{1\text{-}2}$ be one volt, so that $e_{3\text{-}4} = m$ volts. If we now have AB as the input terminals, then we get a figure for overall gain g, such that $e_{3\text{-}4} = m = ge_{\text{in}}$.

Since 1 and 2 are the grid and earthly points of the input circuit, we may let Z_1 be the total grid cathode impedance of the circuit, and by Thevenin's Theorem, if $E_{\text{in}} = \dfrac{em\,Z_2}{Z_1 + Z_2}$ and if $Z_1 = \dfrac{Z_1\,Z_2}{Z_1 + Z_2}$ we get the circuit shown in Fig. 3.

For the easy solution of this circuit, one allowable assumption is necessary, i.e., that Z_3 has negligible loading effect on 3-4. This is allowable, as the choice of Z_3 is free in so far as design considerations will allow.

Since by the transformation giving Z_1, we have made Z_2 infinite, it should be obvious that $\dfrac{1 - e_{\text{in}}}{Z_1} = \dfrac{m - 1}{Z_3}$ or if $Z_3 = nZ_1$, and $g = m/e_{\text{in}}$ we get $g = \dfrac{mn}{1 + n - m}\; n = \dfrac{g\,(m - 1)}{g - m},\; m = \dfrac{g\,(n + 1)}{g + n}$ where obviously g is the overall gain. In these results, any Z, m, n, or g may be complex.

The sign of the real part of g determines whether feedback is negative or positive, and for critical oscillation, $g = \infty$ giving $m = 1 + n$.

If we consider output impedance, we must define our conditions of determining this junction, since there are generally two methods available. If no feedback be used ($n = \infty$), we get identical values whether we determine the load required to give maximum power output, or that required to reduce the terminal potential to half its open-circuit value, by suitable manipulation of theorems, this is obvious for a passive network, but in the case of $n \neq \infty$, we find that the first definition is independent of feedback, since if Z_3 be negligible as a shunt on the load, we have the terminal impedance 3-4 to consider with no reference to feedback. This may be shown as follows:

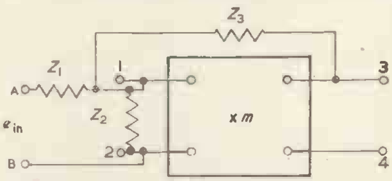


Fig. 2. Amplifier with an external T network

Feedback varies the value of $e_{1\text{-}2}$, but $e_{3\text{-}4}$ being a linear junction of $e_{1\text{-}2}$, then the values of $e_{1\text{-}2}\; e_{3\text{-}4}$ for grid current or cut-off, depend only on the amplifier and not on the feedback circuit. Anyone who has designed an oscillator knows that the valve is treated as a class A amplifier to determine the anode load, after which, feedback is applied to give $g = \infty$.

However if we treat the amplifier as having a potential output across a load, we find that if we determine the load to give half the open-circuit potential, then $Z = Z_o = gD/G_o$ where g_∞ is the value of g determined for an infinite load, and G_0 is the overall mutual conductance of the amplifier at the short-circuited output terminals.

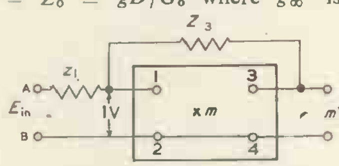


Fig. 3. Development of Fig. 2.

This value is of utmost importance as it is the absolute value of output impedance for any amplifier whatever, and can obviously be determined very closely if, remembering phases, we measure the output potential on open-circuit, and the output current on short-circuit for a constant input.

Referring this to a CV138 (EF91) as a triode connected cathode-follower:—

$$g = m = \frac{-\,G\mu\, R_c}{G\, R_c\,(1 - \mu) - \mu}$$

$G_o = G$, put $G = 7{\cdot}5 \times 10^{-3}$ $\mu = -75$, and we get, if $R_c = \infty$:

$$g = \frac{-\,G\mu}{G(1 - \mu)} = \frac{-\,\mu}{1 - \mu} = 75/76 \text{ and}$$

$$Z_o\, g_\infty / G = \frac{75}{76 \times 7{\cdot}5 \times 10^{-3}} = 131{\cdot}7$$

ohms.

In this case, we get a very close approximation to the correct result by taking $1/G$ as good enough.

Here again, we can differentiate between voltage and current feedback in a generalized statement.

For current feedback

$$g_\infty = m_\infty\; G_o \neq G_f.$$

For voltage feedback

$G_\infty \neq m_\infty\; G_o = G_f$ where G_f is the mutual conductance at the output for one volt of input with the normal load.

By this definition, it is obvious that the cathode-follower is in a state of voltage feedback. If $g_\infty \neq m_\infty\; G_o \neq G_f$, then the feedback is of both types simultaneously.

Although the matter could be developed more fully, I will close by considering a d.c. amplifier, using a CV138(Pentode) with $E_s = 250$, $EG_s = 0$, $E_{HT} = 500$, and a suitable negative bias supply. $g = -20$.

The anode should be centred at 250V and 10mA $R_L = 25\text{k}\Omega$. $E_g = -2$ and $m = -200$. Assume $E_{\text{in}} = +250$V. Then, if E_{in} falls by 5V to 235, E_a rises to 350, and E_g falls to $-2{\cdot}5$V. If the bias supply potential is known, solution for Z_1, Z_2 and Z_3 is possible if any one is chosen arbitrarily. The solution gives $m = 22{\cdot}3$. We now deduce $n_\infty = 22{\cdot}3$, $G_o = G_m = 6{\cdot}5 \times 10^{-3}$, $Z_o = 3{\cdot}43\text{k}\Omega$. If now, we connect this load from anode to h.t. line, then unless we change the bias to give $I_a = 0$, the valve is run outside specification limits. If we connect the same load from anode to cathode, the maximum possible anode potential is 30·5V, and the amplifier is not likely to operate successfully. The gain g however should in either case be reduced to -10. This I think disposes of Mr. Earnshaw's remarks on choice of operating point, and the reason is that the steady state and transient conditions of an amplifier with feedback are not generally identical.

Yours faithfully,

H. MOORBY,

Farnborough, Hants.

A Battery Pre-Amplifier

DEAR SIR,—There is often the need to extend the sensitivity of an audio frequency bridge amplifier by means of a low-level pre-amplifier. A single valve battery-operated unit has been constructed for such a purpose. It is small, self-contained, has a gain of 50 times, and operates for long periods on one set of batteries. For an input resistance of 1MΩ, a signal of 1μV r.m.s. at 50c/s can be detected when the unit is coupled to a tuned amplifier and cathode-ray oscilloscope.

The valve (Fig. A) is a subminiature pentode type DF66, chosen as a compromise between low noise level and battery economy. The filament current of 15mA is supplied from a 1·5V torch cell through a suitable resistor, and the combined anode and screen current of 50μA from a 67·5V portable radio battery. Owing to the small current drain on the batteries, their life is several hundred hours. The valve is designed for wiring directly into a circuit on its stem leads, and in the present application these leads form the sole support. This method of mounting isolates the valve from vibration and mechanical shock. Bias is obtained from the filament dropping resistor, and no grid leak is included, as a d.c. grid return can be provided in the connected circuit. This is desirable since, in some applications, the damping effect of a grid resistor reduces the sensitivity. A resistor of 0·5MΩ is wired in series with the control grid to prevent excessive grid current should a large out-of-balance signal be applied.

The amplifier originally was made as part of the detector of a mains-frequency bridge. In this application shielding from stray electromagnetic fields is extremely important. By far the greatest

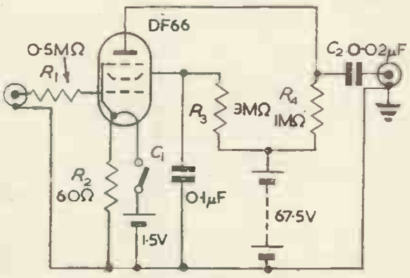


Fig. A. Battery Pre-amplifier circuit R_3 and R_4 = Low noise level type

effect was found to be on the valve, which therefore is screened by three turns of 1in by 0·005in Mumetal strip sprung lightly around the envelope. In addition, the input and output loops are kept as small as possible.

The valve and components are mounted on a lug panel, and are enclosed, with the two batteries, in a 16 s.w.g. aluminium box measuring 5in by 3¾in by 3in, provided with sponge rubber feet. Input and output connexions are made to standard screened connectors.

Yours faithfully,

H. Bairnsfather.

Division of Electrotechnology,
Commonwealth Scientific and Industrial Research Organization,
Sydney, Australia.

Infrasonic Switching

Dear Sir,—It has been suggested that because the optic and acoustic nerves both originate from the fourth cerebral ventricle, they should show analogies. In particular, there should be a phenomenon of persistence of the aural image just as there is a persistence of the visual image.

Experiments to verify this hypothesis have been described by Marro[2] and Montani[1]. In Montani's experiment, sound waves transmitted down a long tube were interrupted by a revolving shutter and the ear was placed at the free end of the tube. Montani claimed that when the shutter rotated at 15c/s, so that the transmitted sound was chopped into equal on and off periods of 0·03sec, the sound received was completely intelligible and the discontinuity imperceptible.

This principle should have applications to two-way communication systems such as loudspeaking telephones. The periodic switching would eliminate unwanted feedback which can lead to instability. Communication systems using infrasonic switching were in fact, proposed by Montani[1]. However, McGuire and Nowacki[3] have reported that repetition of the experiments did not confirm the original claims, but gave no details of their own results. Because of the potential applications of the principle in the field of telephony we have also tried to verify Montani's results.

In view of the intended applications, our experiments were carried out with speech signals transmitted over an electrical circuit which was interrupted, with equal on and off periods. An electrical signal from a microphone was transmitted through an audio-frequency amplifier to a telephone receiver. The circuit between the amplifier and receiver was interrupted by the contacts of a telegraph relay whose coil was energized by the output from a variable-frequency oscillator. One experimenter spoke into the microphone while another listened at the receiver and varied the frequency of the interruptions. When no signal was transmitted, the sound due to the interruptions was negligible. However, it was found impossible to send good quality speech over the circuit although the frequency of interruption was varied between 10c/s and 300c/s during the observations. The speech which was transmitted was intelligible, but a pronounced "warbling" effect was superimposed on it at the frequency of interruption. The effect of this appeared least disturbing when the switching frequency was of the order of 50c/s. At low frequencies (of the order of 15c/s) the length of the interruption periods reduced intelligibility and at high frequencies (of the order of 200c/s) the interruption frequency was itself audible. The most that can be said is that the interrupted speech was intelligible, but its "quality" was so low as to render it completely unsatisfactory for a commercial communication system.

The experimental results which we obtained are consistent with simple modulation theory. The speech signal comprises components at intervals over a wide frequency band. Interrupting the signal at a low frequency produces sidebands of harmonics of the switching frequency about each component of the speech signal. Thus, although the switching frequency may be too low to be audible by itself, its modulation of the speech signal causes many distortion components to be produced at frequencies to which the ear is sensitive. The presence of these modulation products accounts for the warbling effect noticed during the experiments. A similar process takes place, of course, when visual signals are interrupted, but the effect is not noticed because of the very much higher frequency of light waves. Thus, when an audio-frequency wave of say 10^3c/s is interrupted at 50c/s, sideband frequencies are produced whose separation from the carrier is 5 per cent of the carrier frequency. When a light wave of say 6 000Å ($f = 5 \times 10^{14}$c/s) is interrupted at 50c/s the separation of the sidebands frequencies from the carrier is only 10^{-13} of the carrier frequency and no distortion is evident.

Yours faithfully,

J. E. Flood,
R. W. S. Kinsey,

Siemens Brothers & Co. Ltd,
Woolwich,
London, S.E.18.

REFERENCES

1. Montani, A. Infrasonic Switching. *Electronics 19*, 214 (March, 1946).
2. Marro, M. Radiotéléphonie Duplex Simultanée. *L' Onde Electrique 17*, 458 (1938).
3. McGuire, J. H., Nowacki, P. J. A Common-Wave Duplex Pulse-Communication System. *J. Instn. Elect. Engrs. 94*, Part IIIA, 528 (1947)

Relay Scale-of-Two Circuits

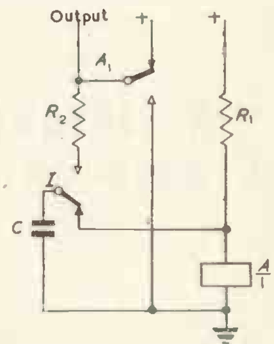
Dear Sir,—The recent article by Mr. R. C. M. Barnes (October and November, 1954 issue) and subsequent letter from Mr. Kendall were most interesting to me as I had recently evolved a circuit similar to Fig. 18 (November issue p. 496) which to my mind has several advantages over the latter.

A current is passed through *A*, such that it will hold on but not operate *A*.

When *I* is operated *C* charges via A_1, R_2 and I-R_2 limiting the current. Upon release of *I*, *C* is discharged through *A* thus operating it, the standing current holding *A* in this position.

When *I* is operated a second time *C* is discharged via A_1—*A* being operated—and upon release of *I*, *C* effectively short-circuits *A* thus releasing it.

Relay scale-of-two circuit



The rate of operation is mainly limited by the charging rate. On the other hand R_2 may be increased deliberately so that the circuit will respond only to pulses of more than a pre-determined length.

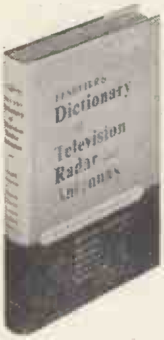
The circuit has not been tried at more than some 300 cycles per min. *C* being 16μF R_2 100Ω and *A* a 3 400Ω high speed relay.

One advantage of this circuit is that the power requirements are identical in both states which simplifies power supply design, and surges on the supply line do not affect *A* as R_1/C form an effective filter. The circuit is simple and uses no special components and works on pulses of any duration.

Yours faithfully,

R. G. Wicker.

Coundon, Coventry.



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BOOK REVIEWS

Applied Electronics

By T. S. Gray, 881 pp. 364 figs. Demy 8vo. J. Wiley, Inc., New York. Chapman & Hall Ltd., London. 1954. Price 72s.

SO many books have been written round the subject of radio and electronics, especially in America, that a reviewer is reaching the stage where he has to look for fine points of detail in order to distinguish one book from another. This is a work sponsored by M.I.T. and is a second edition so that it merits careful consideration. One pleasant feature (regrettably rather rare in American publications) is the author's recognition of the work that has been done outside America and there are many references to European authors; the bibliography shows to a less degree the same catholicity. Another excellent feature is the answers provided to the numerical problems (a good characteristic of American student books) given at the end of each chapter.

The book certainly differs from others on the same theme in that its first five chapters (out of 13) concentrate on the physics of the valve. This is done quite well and should appeal to the physics or engineering student who is going to concern himself with the development of devices depending on electron action in vacuum, gas or vapour. Electron ballistics, emission from heated metals, conducting through a vacuum, gas and vapour are all treated and there is a short reference to semi-conductors.

In the next six chapters the author follows well trodden paths going from rectifiers to amplifiers and from oscillators to demodulators. The approach is academic rather than practical, and from this point of view the title is a little misleading; "Principles of Applied Electronics" would be more appropriate. The author is strictly logical in his theoretical development. Thus we find in the section on the Miller Effect that input capacitance is

$$C_{gk}+C_{ga}(1-A)$$

where A is the algebraic gain from output to input and due to the phase change in the valve it has a negative real component. The same is true of negative feedback. In the cathode-follower the expression becomes

$$C_{gk}/1+A+C_{ga}$$

where A has now a positive real component and is the stage gain from grid-cathode voltage to output. Many will be more familiar with the expression $C_{gk}(1-A')$ where A' is the stage gain from input (grid-earth) to output (cathode-earth). The vectorial presentation of valve voltages which puts μE_g and E_g in phase is preferred so that the output voltage vector is 180° out of phase with the product of the current and output load impedance. This may be a little bewildering at first to the reader who has some knowledge of electronics but will present no problems to the student who is being guided for the first time through the intricacies of the subject by the author.

In the section on rectifiers with a capacitance load some mention should have been made to the work (used by the author in his theoretical development) of Marique, which preceded Schade's publication by about eight years (*Wireless Engineer*, January 1935, p.17).

The last chapter is a useful summary of the characteristics and applications of transistors.

This is a good sound book, painstakingly developed; it is likely to have a sale in America among physics or engineering students specializing in electronics. It seems improbable that it will be a serious competitor with books (American as well as English) which have already established themselves in the English market.

K. R. STURLEY

Astronomical Photoelectric Photometry

Edited by Frank Bradshaw Wood. 141 pp., 39 figs. Royal 8vo. American Association for the Advancement of Science, Washington. 1953. Price $5.

THE advent of the photomultiplier has played an important part in the development of astronomical photoelectric photometry. In the early days of the subject photocells and d.c. amplifiers were the standard equipment, but in the last decade many new electronic techniques of measurement have been applied to this fascinating problem, and the investigator has now a wide choice of tools and methods. The best method to use in a given set of circumstances, e.g. d.c., a.c., or pulse will depend, no doubt, on the experience of the investigator and the equipment available to him, but the present volume should be of considerable assistance to him in those cases where a choice of techniques or methods arises.

The volume comprises a collection of papers presented, in a somewhat abbreviated form, at a symposium held in Philadelphia on December 31st, 1951, by Section D (Astronomy) of the American Association for the Advancement of Science. The papers are entirely concerned with problems of instrumentation and technique. Nine papers are presented. The first three deal with direct-current, alternating-current and pulse techniques respectively and each provide a good introduction to the subject. In some ways, however, the first paper is a disappointment, and the reviewer would have preferred a considerable cut in bibliography (575 references are quoted in the first paper) and the extra space used to provide a more detailed treatment of the d.c. techniques of measurement.

The fourth and fifth papers are very short contributions describing particular pulse counting installations, while the sixth paper gives a very complete and interesting account of the performance of two pulse-counting stellar photometers used at the Observatory at Cambridge, England. The seventh paper describes the use of servomechanisms in the photo-electric photometry of stars, while the last paper considers the limits of sensitivity and precision attainable by photo-electric methods. This chapter is the best written and the most interesting in the whole book. To the newcomer to the subject this chapter is an excellent introduction and puts many of the statements in the earlier chapters in their right perspective. The reviewer's feeling is, in fact, that this chapter could logically have been the first and not the last.

The book, like many others giving a collection of papers all written by different authors suffers from the disadvantage of too much repetition and not sufficient explanation of some of the details. It is nevertheless a valuable book for the astronomer and will be found interesting reading to the electronic specialist.

DENIS TAYLOR.

Foundations of Potential Theory

By Oliver D. Kellogg. 384 pp. 30 figs. Demy 8vo. Dover Publications. Inc., New York. 1954. Price $1.90 (paper), $3.95 (cloth).

THIS book was originally published in English some twenty-five years ago in the Springer yellow series "Grundlehren der mathematischen Wissenschaften" and has now been re-issued by Dover Publications. The revival of such an excellent treatise on the subject is to be welcomed, even if this is not a new edition, so that references and bibliography have not been brought up to date. The subject matter is dealt with from an advanced mathematical point of view, particularly so in the second half of the book. However, the physical basis and applications of potential theory to different fields of physics (gravitation, electrostatics, hydrodynamics) are never lost sight of. One of the most valuable contributions of the book lies in its rigorous treatment of mathematical proofs, especially the full discussion of existence theorems. But this should not be taken to imply that it is only meant for the pure mathematician. A much wider class of readers will find a great deal to learn from it. They will be helped in this by large numbers of examples.

G. FIELD

Atomic Energy and its Applications

By J. M. A. Lenihan. 265 pp., 70 figs. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1954. Price 22s. 6d.

THIS book is written for the reader of modest scientific attainments who wants to know more about the achievements and potentialities of nuclear science. It is neither a textbook of atomic physics nor a catalogue of its applications, but combines the most valuable features of both. Written in a simple, straightforward style, this book will be of interest to teachers and students.

Neuartige Aluminiumleiter in Starkstromfreileitungen (Novel Aluminium Conductors in Overhead Power Lines)

By Dr. Milan Vidmar. 208 pp. 29 figs. Demy 8vo. Academia Scientiarum et Artium Slovenica, Institution Oeconomiae Electricae, Ljubljana. 1953.

THIS monograph is of greater interest to the power engineer than to the electronic engineer. But as a preceding book by the same author dealing with a similar subject was reviewed in the August 1953 issue, only a brief account of its principal contents shall be given.

The author's proposal to use pure aluminium conductors not reinforced by steel cores for high power transmission is given a thorough theoretical foundation. The eight chapters of the book deal with the following subjects: the mechanical states of overhead power lines; the three terms of the equation of state; the additional mechanical loads; the largest sag during summer; the steel-aluminium conductor; sags and span; complete equivalence of different kinds of conductors; concluding considerations. A brief list of references is given containing mainly the author's own publications. A slight misprint was noticed; Matkin's paper appeared in 1952, not in 1925.

R. NEUMANN.

The Inventor of The Valve

By J. T. MacGregor-Morris. 134 pp. Demy 8vo. The Television Society. 1954. Price 10s.

COINCIDING with the jubilee of the invention of the thermionic valve, the Television Society published this short biography of Sir Ambrose Fleming, written by his former student and assistant, Professor J. T. MacGregor-Morris, with a foreword by Professor E. W. Marchant and an appendix of personal recollections by Mr. Arthur Blok. The text is illustrated with reproductions of original notes and letters, many of which are published for the first time.

The edition is limited to 1 000 copies, and the book will thus not only be a unique record of Fleming's achievements but will increase in value as time goes on.

FBI Register 1955

1 089 pp. Crown 4to. 27th Edition. Iliffe & Sons Ltd. 1954. Price 42s.

AN important new feature of this edition is the French, German and Spanish Glossaries. Each glossary gives a translation of every heading used in the buyers' guide, the headings being numbered for easy reference between the translations and the main section.

There are eight sections in the new Register, each one having a reinforced index card for ease of reference. These include a buyers' guide which classifies over 6 800 FBI member firms under more than 5 000 trade headings, and a comprehensive alphabetical directory of all these firms. In other sections will be found useful information about trade associations, proprietary names, trade marks, etc.

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ELECTRONIC EQUIPMENT

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

Temperature Control Unit

(Illustrated below)

THE control unit type N856 is designed for the accurate control of the temperature of platens, ovens, rooms, containers, liquids, gases, etc., in the range −96°C to 428°C.

It normally utilizes a flat mica resistance type sensitive element which is mounted in close proximity to the medium to be controlled, but for immersion in liquids, use in corrosive or damp atmospheres, or for controlling temperatures above 300°C, a sealed platinum bulb thermometer may be employed as the temperature sensitive element.

The element forms one arm of a

bridge circuit, another arm of which is made variable to provide the adjustment of temperature. The bridge out-of-balance voltage is amplified and used to operate two separate relays at slightly different temperatures, the temperature difference being adjustable between 0 and 3°C.

This arrangement enables the three conditions "temperature low," "temperature correct" and "temperature high" to be detected, and the unit can therefore be used to control not only electrically heated elements, but also heating or cooling processes using motor controlled valves, etc.

The sensitivity of the equipment is such that the relays operate and release on changes of the control resistance of 0·1 per cent. This represents a temperature change of 0·25°C for the mica resistance element and 0·4°C for the platinum bulb thermometer at 100°C.

Airmec Ltd,
High Wycombe,
Buckinghamshire.

Pulse Generator

(Illustrated above right)

THE pulse generator type R100A provides a wide variety of pulses for the checking and calibration of ratemeters, scalers, pulse amplifiers, etc.

The pulse generator provides positive and negative, single or double rectangular pulses with adjustable spacing in two separate channels, A and B.

The A and B output controls are both calibrated 0 to 50 volts, B output pulses are variable in width.

The delay control is calibrated with three ranges, 0 to 20μsec, 20 to 220μsec and 220 to 1 000μsec.

Panax Equipment Ltd,
173, London Road,
Mitcham,
Surrey.

Glass Fibre Laminates

A NEW range of glass fibrepolyester resin laminates is now available. Manufactured under the trade name of "Rilglaze" they are available in various grades and have high mechanical strength, good heat resistance and insulation properties.

A general property grade ("S") is available which is generally used in the form of blanks in electrical components. This grade has good punching and machining properties.

Rubber Improvement Ltd,
Rilex Works,
Wellingborough,
Northamptonshire.

Oscillograph Tube

(Illustrated below)

A NEW Emitron oscillograph tube is now available. This tube, the type 4EP1, employs post-deflexion acceleration which gives the advantage of high deflexion sensitivity combined with good brightness: the sensitivity being about 1mm/V with a final accelerator potential of 2kV. For high writing speeds this voltage can be increased up to 8kV.

The screen diameter is 4in and of medium persistence. Side connexions are used for the deflector plates to allow the tube to operate at very high frequencies.

Electronic Tubes Ltd,
Kingsmead Works,
High Wycombe,
Buckinghamshire.

R.F. Induction Heater

(Illustrated below)

THE induction heater type EH.120 has versatile output coupling arrangements which enable it to be used for a wide variety of heating processes.

R.F. power output is 12kW for continuous, or 14kW for intermittent working (50 per cent duty cycle) into work of high power factor, or 500kVA and 700kVA respectively into work of low

power factor or where the coupling efficiency is low. A hand wheel on the front control panel provides a smooth and continuous control of the r.f. voltage supplied to the output terminals.

Work-matching capacitors in varying combinations up to 0·14μF are incorporated, but in cases where r.f. power is fed to a remote work position via a transmission line, the capacitor can be readily omitted. The continuously variable output coupling system is very suitable for feeding into a coaxial line, and a dual-range line ammeter on the control panel facilitates accurate matching to a transmission line and external work circuit.

For accurate timing of operations, a timing unit with built-in hour meter is incorporated. The normal range is 0 to 30 seconds, but other ranges from 0 to 10 seconds up to 0 to 4 hours are available to order.

E.M.I. Factories Ltd,
Hayes,
Middlesex.

Ceramic Capacitors

A NEW range of inexpensive ceramic capacitors, known as "Cascaps," has been introduced by Plessey.

The material used is of the barium titanate type, having high permittivity and dielectric strength, enabling miniature capacitors to be produced, for operating at high voltages over a wide temperature range and used in applications where precision of capacitance is not of primary importance.

"Cascaps" are available in two basic types, one for purposes such as r.f. decoupling and the other for radio interference suppressors.

The former types are constructed in capacitances ranging from 5·00pF to 0·01μF, 500V d.c. or 300V a.c. working with a breakdown voltage of 4 000V d.c. and an insulation resistance of $10^{11}\Omega$. A 0·005μF capacitor measures only 0·664in in diameter and is 0·187in thick.

All "Cascap" capacitors are finished in a recently developed paint coating, which forms a thick protective layer on the disk and provides additional mechanical strength. This coating has exceptionally high insulation resistance, while it also prevents entry of moisture, so avoiding the necessity for wax impregnation.

The Plessey Co. Ltd,
Ilford,
Essex.

Industrial Servo Unit

THE industrial servo unit type R.1219 is designed to help particularly industrial laboratories which are experimenting with servo mechanisms for various control purposes.

The unit provides all power supplies for operating the servo motor, and is suitable for rack mounting or for bench use.

When used in conjunction with a servo motor such as the Evershed & Vignoles' type F.A.2, the unit can be used for many purposes such as, high power voltage stabilization, controlled speed drive, follower systems, process control.

An input signal of 0·2V will result in the maximum d.c. field current being applied to the motor (80mA centre-tapped). The armature current is 1A d.c.

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

Permanent Soldering Bits

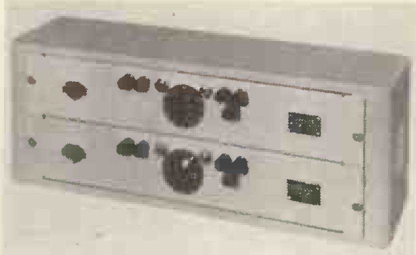
THE manufacturers of Litesold soldering irons claim to have evolved a permanent bit for soldering irons. It is registered under the trade name "Permabit." It is claimed that this bit lasts indefinitely, does not become pitted or lose its face, and requires no reshaping, filing or maintenance. These Permabits are available in a fixed bit range of instruments and also as bits for replaceable types in all sizes.

Light Soldering Developments Ltd,
106, George Street,
Croydon,
Surrey.

Filter Cabinet

(Illustrated above right)

MULLARD high- and low-pass filters are now available in pairs in a neat bench-type housing. These filters, based on a design originated by the British Post Office, have ten cut-off frequencies between 400c/s and 17kc/s. They may be used separately as a high-pass or low-pass filter, or connected in series to form



a band-pass filter of variable bandwidth.

The filters are mounted on standard 19in panels only 3½in high. Their compactness results from the use of Ferroxcube core material for the inductors. This makes it possible to construct filters having a high image impedance, thus enabling compact high stability silvered mica capacitors to be used. Matching transformers reduce this high impedance to 600Ω.

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

Key Operated Switches

(Illustrated above)

CRATER Products announce that any of their existing range of rotary switch can now be supplied fitted with a barrel type key operated mechanism. They point out that this should be particularly useful in cases where it is desired to prevent unauthorized tampering with equipment, as for instance, in some industrial installations.

Crater Products Ltd,
The Lye, St. Johns,
Woking,
Surrey.

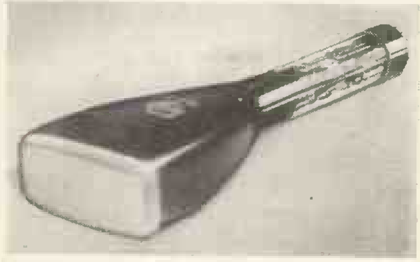
Superspeed Soldering Iron

(Illustrated below)

THIS new soldering iron weighs only 3½oz and heats up to soldering temperature in less than 6sec. It may be

used on any supply voltage between 2·5 and 6·3V; a 4V transformer is normally supplied. It is switched on by light pressure on the switch ring, and is automatically switched off when not in use. It is claimed that, due to the principle upon which the iron works, it is more powerful than a conventional 150W iron.

Enthoven Solders Ltd,
89, Upper Thames Street,
London, E.C.4.



Rectangular Faced Oscilloscope Tube

(Illustrated above)

IN radar and oscilloscope displays, it often happens that only a small horizontal strip in the centre of the screen of a cathode-ray tube is occupied by the trace. If instead of a conventional circular face a face of rectangular form is used, the trace can occupy the whole of the working screen area. Such a rectangular screen has been adopted in the new Mullard DG16-21 tube, which enables much equipment space to be saved.

The DG16-21 tube has a screen size of 5¼in by 1½in. The deflexion sensitivity is of the order of 0·2mm/V. The angle alignment between X and Y plates is kept within one degree of the nominal value of 90 degrees: this close tolerance ensures the high degree of perpendicularity necessary where accurate measurements have to be made. The tube has a 6·3V heater, and would normally operate at a final anode voltage of 6KV.

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

Push-Button Switch

(Illustrated above)

THIS new addition to the Bulgin range of instrument pushes has single ½in diameter hole fixing. The switch has a single pole push for on action. The brass body bush is chromium plated and the rear fixing butt is nickeled. It is rated at 1A at 110V a.c. The list number is M.P.12.

A. F. Bulgin & Co. Ltd,
Bye Pass Road,
Barking,
Essex.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 30 March. Time: 6.30. p.m.
Held at: The London School of Hygiene and Tropical Medicine, Gower Street, W.C.1.
Discussion: The Maintainability of Service Equipment.

West Midlands Section

Date: 9 March. Time: 7.15. p.m.
Held at: Wolverhampton and Staffordshire Technical College, Wulfruna Street. Wolverhampton.
Lecture: Electrical Standards in Electronics.
By: P. M. Clifford.

North-Eastern Section

Date: 9 March. Time: 6 p.m.
Held at: Neville Hall, Westgate Road, Newcastle-upon-Tyne.
Lecture: The Application of Negative Feedback to Electrical Measuring Instruments.
By: F. J. U. Ritson.

Merseyside Section

Date: 3 March. Time: 7 p.m.
Held at: College of Technology, Byrom Street, Liverpool 3.
Symposium on Electronics in Industry.

North-Western Section

Date: 3 March. Time: 7 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester.
Lecture: Computing Circuits in Flight Simulators.
By: A. E. Cutler.
Date: 31 March. (Time and place as above.)
Lecture: Industrial Applications of Electrical Control.
By: J. A. Sargrove.
Followed by Annual General Meeting.

Scottish Section

Date: 10 March Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders. Elmbank Crescent, Glasgow.
Lecture: Computing Circuits in Flight Simulators.
By: G. B. Ringham and A. E. Cutler.

BRITISH SOUND RECORDING ASSOCIATION

Date: 25 March. Time: 7 p.m.
Held at: The Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2.
Lecture: High Fidelity Reproduction in the home using the Metal Cone Loudspeaker.
By: F. H. Brittain.

THE INSTITUTION OF ELECTRICAL ENGINEERS

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

Radio Section

Date: 2 March.
Lectures: Some Comparative Directional Measurements on Short Radio Waves over different transmission Paths.
By: E. N. Bramley.
Some Aspects of the Rapid Directional Fluctuations of Short Radio Waves reflected at the Ionosphere.
By: E. N. Bramley.
The Rapidity of Fluctuations of Continuous Wave Radio Bearings at High Frequencies.
By: W. C. Bain.
Sources of Error in U-Adcock High-Frequency Direction Finding.
By: K. C. Bowen.
Date: 9 March.
Lecture: Artificial Reverberation.
By: P. E. Axon, C. L. S. Gilford and D. E. L. Shorter.

Measurements Section

Date: 15 March.
Lectures: The Mechanism of Sub-harmonic Generation in a Feedback System.
By: J. C. West and J. L. Douce.
The Transient Behaviour of Remote-Position Control Systems with a Hard Spring Non-Linear Characteristic.
By: J. C. West and P. N. Nikiforuk.

East Midland Centre

Date: 15 March Time 6.30. p.m.
Held at: The East Midland Electricity Board, Service Centre, Derby.
Lecture: The Possibilities of a Cross-Channel Power Link Between the British and French Supply Systems.
By: D. P. Sayers, M. E. Laborde and F. J. Lane

Cambridge Radio Group

Date 15 March. Time: 6 p.m.
Held at: The Cambridgeshire Technical College.
Lecture: A Caesium Clock.
By: J. G. Yates.

North-Eastern Radio and Measurements Group

Date: 7 March. Time 6.15. p.m.
Held at: King's College, Newcastle-upon-Tyne.
Lectures: Standard Frequency Transmissions.
By: L Essen.
The Standard Frequency Monitor at the National Physical Laboratory.
By: J. McA. Steele.
Standard Frequency Transmission Equipment at Rugby Radio Station.
By: H. B. Law.

North Midland Centre

Date: 15 March. Time: 6.30, p.m.
Held at: The Technical College, Bradford.
Lecture: Colour Television.
By: G. N. Patchett.

North-Western Measurements Group

Date: 22 March. Time: 6.15. p.m.
Held at: The Engineers' Club, Albert Square, Manchester.
Discussions: The Servicing of Electronic Measuring Instruments and its Effects on their Design.
Opened by: Denis Taylor.

South-East Scotland Sub-Centre

Date: 15 March. Time: 6.30. p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lectures: The Genesis of the Thermionic Valve.
By: G. W. O. Howe.
Thermionic Devices from the Development of the Triode up to 1939.
By: Sir Edward Appleton.
Developments in Thermionic Devices since 1939.
By: J. Thomson.

North Staffordshire Sub-Centre

Date: 11 March. Time: 7 p.m.
Held at: The Electricity Showrooms, Stoke-on-Trent.
Lecture: The Possibilities of a Cross-Channel Power Link between the British and French Supply Systems.
By: D. P. Sayers. M. E. Laborde and F. J. Lane.

Rugby Sub-Centre

Date: 2 March. Time: 6.30 p.m.
Held at: The Rugby College of Technology and Arts.
Lecture: Developments in Thermionic Devices since 1939.
By: J. Thomson.

Southern Centre

Date: 9 March. Time: 6.30 p.m.
Held at: Portsmouth College of Technology.
Lecture: The Manchester-Kirk o'Shotts Television Radio Relay System.
By: G. Dawson, L. L. Hall, K. G. Hodgson, R. A.Meers and J. H. H. Merriman.
Date: 16 March. Time: 7.30. p.m.
Held at: R.A.E. Technical College, Farnborough.
Lecture: A Transatlantic Telephone Cable.
By: M. J. Kelly. Sir Gordon Radley, G. W. Gilaman and R. J. Halsey.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 8 March. Time: 5 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2.
Lecture: Some Non-Linear Magnetic Devices and their Application to Telecommunications.
By: J. W. McPherson.

THE INSTITUTE OF PHYSICS

Date: 24 March. Time: 6.30 p.m.
Held at: The Institute's House, 47, Belgrave Square, London, S.W.1.
Lecture: Recent Developments in Ultrasonics.
By: E. G. Richardson.

Midland Branch

Date: 17 March. Time: 5.15, p.m.
Held at: The Physics Department, University of Birmingham.
Lecture: Electronic Musical Instruments.
By: K. A. Macfadyen.

THE TELEVISION SOCIETY

Date: 10 March. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: Distributed Amplifiers.
By: W. S. Percival.

PUBLICATIONS RECEIVED

NRDC BULLETIN is a quarterly review of inventions available for introduction to industry, published by the National Research Development Corporation, 1 Tilney Street, London, W.1.

A REVIEW OF PRODUCTIVITY IN THE DIESEL LOCOMOTIVE INDUSTRY is the fifteenth in a series dealing with productivity in various industries. Its main purpose is to examine the steps taken by diesel locomotive manufacturers in recent years towards a fuller use of their resources of equipment and labour. The British Productivity Council, 21, Tothill Street, London, S.W.1. Price 2s.

BOWATERS BUILD A MILL IN TENNESSEE is the latest chapter in the story of an expanding organization that serves five continents. The Bowater Paper Corporation Ltd., Stratton Street, London, W.1.

AUTOMATIC PROTECTION OF A C. CIRCUITS by G. W. Stubbings is the 4th edition of a book first published in 1934. It has been completely revised, and the chapters dealing with relays and protective systems and the glossary of protective gear engineering terms have been re-edited. Chapman & Hall Ltd., 37 Essex Street, Strand, London, W.C.2. Price 50s.

ZONAX, FERRAX, KLEENAX, KLENEWELL and UNIMAX are the titles of five pamphlets covering the comprehensive range of Canning metal cleaners developed specifically for the electroplating and metal finishing industries. These cleaners are intended for the removal of grease or oil left on surfaces after machining, stamping, spinning, pressing or polishing. W. Canning & Co Ltd, Great Hampton Street, Birmingham, 10.

A SERVICE TO INDUSTRY is a booklet which briefly describes the organization and work at the factory of C. Robinson & Partners Ltd at Cheadle, Cheshire, which produces electronic controls and instruments. C. Robinson & Partners Ltd, 287 Deansgate, Manchester, 3.

ALKYD MOULDING COMPOUNDS are a selection of leaflets describing the Plaskon Alkyd moulding compounds which Resinuous Chemicals Ltd. are manufacturing in this country under licence from the Barrett Division, Allied Chemicals and Dye Corporation, Toledo, Ohio. Resinous Chemicals Ltd., Blaydon, County Durham.

NIMONIC 90 FOR SPRINGS is a recent data sheet published by Henry Wiggin & Co Ltd and presents, in convenient form, tabular data for the designer. The Nimonic series of high temperature alloys have, for many years, been the standard materials for moving blades in gas turbines. Their use in other fields, however, is rapidly extending where high strength at high temperatures is required. Typical of such applications is the use of Nimonic 90 for springs operating at extreme temperatures. Copies of this data sheet are available from the Publicity Department, Henry Wiggin & Co Ltd, Thames House, Millbank, London, S.W.1.

PRACTICAL LESSONS IN METAL TURNING AND SCREWCUTTING by Percival Marshall is a revised edition of the original published some five years ago and is mainly intended as a guide to the methods of using lathes of the kind usually found in the workshops of general engineers, cycle and motor repairers, instrument makers, model makers, and amateurs. Percival Marshall & Co. Ltd., 23 Great Queen Street, London, W.C.2. Price 3s. 6d.

PICTURE BOOK OF TELEVISION TROUBLES is the first in a series of volumes which reports the results of fault finding in receivers in the John F. Rider Laboratories, New York. John F. Rider Publisher, Inc., 480 Canal Street, New York 13. Price $1.35.