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BRITISH INSTITUTION OF RADIO ENGINEERS

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"To promote the advancement of radio, electronics and kindred subjects by the exchange of information in these branches of engineering."

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THE APPLICATIONS OF ELECTRONICS

MANY of the Institution's meetings being held during the current session deal with applications of radio technique to fields other than communications—a process generally described as "electronics". Many of these new developments have in fact originated through the work of the professional radio engineer and have been discussed within the Institution for the past three decades. Professional engineers in other fields have adopted electronic techniques to solve their own particular problems, and this has led to increasing co-operation with radio engineers.

Industrial electronics is the subject of a three-day symposium to be held in London at the beginning of April. The Institution's Conventions in 1954 and 1957 dealt with many aspects, notably the application of the techniques of automation to a wide range of industries. The basic theory of the control of production processes by electronic means has its roots in the work done during the twenties on feedback amplifiers for communications purposes; it is only during the past decade, however, that the greatly increased force of radio engineers built up during the war to develop radar and communications has been able to convince industry generally of the great contribution which electronics can make to increasing productivity and efficiency. Measurement, control and computation are techniques in which the use of electronic devices excels and these will receive due consideration in the forthcoming symposium which will deal both with basic devices and with applications.

The application of electronics to medicine was for many years the aim of a small group of enthusiastic radio engineers. Acceptance of their ideas by the medical profession was slow, but during the last few years there has been much closer liaison between doctor and engineer, and this has spread on an international basis. An Institution symposium being held this month (March) will deal with an especially fascinating and worthwhile aspect, that of aiding the handicapped—the blind, the deaf and dumb, and the crippled. Although some of these aids are in a very early stage and call for the surmounting of formidable problems, this meeting will fulfil a useful purpose if it only starts radio engineers thinking of ways to solve some of the problems.

A third symposium, to be held in Birmingham in July, will be on sonar systems. While the programme will be devoted mainly to various communication problems, sonar has its uses in fishing, navigation and oceanography, and it is interesting to note that rather similar techniques are employed in a new guidance device for the blind which will be described during the March symposium.

This last point sums up the true value of Institution meetings—in brief, cross-fertilization. Interest in other sciences and technologies is invariably rewarding; it often leads to new approaches to common problems. It is for this reason that the Institution always welcomes opportunity for joint meetings and association with other bodies.

INSTITUTION NOTICES

I.E.E.–Brit.I.R.E. Collaboration

The Councils of the Institution of Electrical Engineers and The British Institution of Radio Engineers have set up a Joint Committee with the following terms of reference:

"To suggest and to examine means by which, through collaboration between the two Institutions, the progress of electronic and radio engineering can be fostered, and to make recommendations for submission to the Councils of the two Institutions."

The following represent the two Institutions on the Committee:

Brit.I.R.E.

Mr. L. H. Bedford, C.B.E., M.A., B.Sc., F.C.G.I., M.Brit.I.R.E.

Mr. I. Maddock, O.B.E., B.Sc.(Hons.), M.Brit.I.R.E.

Mr. W. E. Miller, M.A.(Cantab.), M.Brit.I.R.E.

Colonel G. W. Raby, C.B.E., M.Brit.I.R.E.

I.E.E.

Sir Harold Bishop, C.B.E., B.Sc.(Eng.), F.C.G.I., M.I.E.E.

Mr. R. J. Halsey, C.M.G., B.Sc.(Eng.), F.C.G.I., M.I.E.E.

Sir Hamish D. McLaren, K.B.E., C.B., D.F.C.*, LL.D., B.Sc., M.I.E.E. (*Chairman*).

Mr. T. B. D. Terroni, B.Sc., M.I.E.E.

The first meeting of the Joint Committee on 6th February 1962 dealt with the possible participation of the Institution in the Television Conference which the Institution of Electrical Engineers is organizing and which will be held from 31st May to 7th June 1962.

Further information and advice as to how Brit.I.R.E. members may register are given on page 194.

Symposium on Sonar Systems

The Institution, through its Electro-Acoustics and Radar and Navigational Aids Groups, is sponsoring a two-day meeting on Sonar Systems to be held in the Electrical Engineering Department of Birmingham University on 9th–10th July 1962. The Acoustics Group of the Institute of Physics and the Physical Society is collaborating in the Symposium which will include contributions from overseas.

Papers are invited on the following subjects:

Transducers, Arrays and Scanning Systems (including Directional Systems in general)

Detection and Display

Aspects of Propagation relevant to System Performance (e.g. Echo Structure, Target Characteristics, Fluctuations).

Offers of papers, which should preferably be accompanied by a 150-word synopsis, should be sent to the Brit.I.R.E., 9 Bedford Square, London, W.C.1.

Commonwealth Conference on Communications Satellites

On 2nd March the Postmaster-General, the Rt. Hon. J. R. Bevins, M.P., announced that "... Commonwealth Governments have welcomed a proposal to arrange a special conference to discuss the problems associated with the use of satellites and their potentialities for Commonwealth communications. Representatives of Commonwealth Administrations will be meeting in London at the end of this month to prepare a report for consideration by Governments."

The conference will last two to three weeks and the countries which have accepted invitations are: Australia, Canada, Ceylon, Ghana, India, New Zealand, Nigeria, Pakistan, the Federation of Rhodesia and Nyasaland, and Sierra Leone. The Director General of the British Post Office, Sir Ronald German, C.M.G., will be the chairman, and Sir Robert Harvey, K.B.E., C.B., will be the leader of the U.K. delegation. The secretary to the Conference will be Mr. W. Stubbs, C.B.E., M.C. (Member), who is Secretary-General of the Commonwealth Telecommunications Board.

The President, Officers and Council of the Institution are holding a reception at the Savoy Hotel, London, on 29th March to welcome the delegates.

The 1963 Convention—Preliminary Notice

As stated in the February *Journal*, the subject of the 1963 Convention will be coupled with National Productivity Year and will cover the general theme of Electronic Aids to Productivity. For the first time since the 1951 Convention the venue will be the University of Southampton. Residential accommodation will be in the University's Halls of Residence. The Convention will start on Tuesday, 16th April, and disperse on Saturday, 20th April. A Convention Committee is being appointed by the Council and further details will appear in the *Journal* in the near future. Meanwhile offers or suggestions of papers relevant to the theme are invited.

Special General Meeting of the Institution

A Special General Meeting of Corporate Members of the Institution will be held on Wednesday, 2nd May 1962 at 5.30 p.m. in the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1. The motion before the meeting will be the adoption of the Bye-Laws, as is required in the Charter of Incorporation.

The meeting will be followed at 6 p.m. by a paper sponsored by the Electro-Acoustics Group Committee on "Loudspeaking Telephones".

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Lightning—Facts and Fancies

By

W. A. GAMBLING, Ph.D., B.Sc.

(Associate Member)†

The Chairman's Address to the Southern Centre of the Institution given in Farnborough, Hampshire on 26th September 1961.

Summary: The lightning stroke, with its accompanying clap of thunder, has always been regarded with awe as one of nature's more impressive manifestations. Not surprisingly this has given rise to a considerable amount of mythology on the subject, some of which persists even today. After a discussion of these "fancies" the paper turns to the work of Benjamin Franklin who made the first serious study of lightning, resulting in the invention of the lightning rod. Unfortunately this gave rise to the famous controversy concerning sharp and blunt conductors. An outline of current theories of the lightning discharge is given and the paper ends with some comments on the effects of thunderstorms, particularly those which are of interest to the radio engineer.

Hell itself may be contained
within the compass of a spark. THOREAU

1. Some Early Beliefs

The lightning flash, with its thunderous accompaniment, is a very impressive natural phenomenon particularly when it occurs in close proximity to the observer. It is perhaps not surprising therefore that the ancients, who were somewhat less scientifically minded and consequently more prone to superstition than we are today, should have built up a considerable mythology on the subject in the belief that the terrible and unpredictable power of the thunderbolt was a manifestation of the wrath of divine beings. We must not forget however that before the invention of the lightning arrester the destruction wrought by thunderstorms was considerably greater than it is today.

The Greeks, for example, believed[1] that Zeus often intervened with his thunderbolts in battles which his favoured side, or the side which produced the more effective prayers and sacrifices, was likely to lose. It was, of course, very unwise to antagonize the gods and Virgil records that a certain Prince, wishing to be deified, drove his chariot over a metal bridge to imitate thunder while throwing torches to serve as lightning. This was a very foolhardy thing to do. For his temerity Zeus hurled a real thunderbolt at him and burnt him up.

One of the Romans' many beliefs[1] was that persons killed by lightning had incurred the wrath of the gods so that they were buried hurriedly and without ceremony at the spot where they were struck. Animals killed were considered unclean and the flesh was not eaten. It is also on record that one Roman citizen, who was impious enough to rebuild his house after it had been destroyed in a thunderstorm, was ordered by the authorities to take it down again, presumably because it was obvious that the Gods did not wish it to be there. The Emperor Tiberius, who was very much afraid of thunder, used always to put on a crown of laurel during thunderstorms in the belief that this afforded complete protection.

In fact lightning even figured very prominently in the affairs of state in ancient Rome,[1] for there existed an eminent body known as the College of Augurs. It was the duty of an augur to ascertain the views of the gods on important matters. This he did by observing the three portents of lightning, birds, and shooting stars. If, in facing south, the lightning was observed to pass from left to right the omens were favourable, but when it passed in the reverse direction the augur would report that Jupiter did not approve of the proceedings in the Forum or other public meeting and these were therefore cancelled. The Augurs were thus very powerfully placed since no independent confirmation of their observations was necessary and in 59 B.C. the entire legislative programme of Julius Caesar was held up by one augur in this way. Almost the only semblance of a scientific inquiry at this time was made by Lucretius whose thoughts on the nature of the universe make fascinating reading.[2] He asked why Jupiter should be so wasteful in hurling his thunderbolts on the sea, mountains, trees, and his own temples, and why he had to wait for thick clouds before launching them. He expressed doubts that even Jupiter could aim thunderbolts simultaneously in several widely separated places.

The Norsemen thought that lightning was caused by their god Thor hurling his magic hammer to the ground. As evidence they sometimes found in the earth unusual pieces of stone or iron which were

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assumed to be pieces broken from the hammer by the violence of the impact. Finds of this kind are made even today but we know them as stone-age implements and meteorites.

Some primitive tribes still believe that a strange thunder-bird which inhabits the clouds dives to earth, the vivid wings appearing as lightning and the loud beating of the wings causing the thunder. They point to the marks of its claws on the trees and huts which have been struck, as confirmation.

2. The Effects of Lightning

Before the advent of the lightning rod severe damage and many deaths were caused by lightning. In Europe it was the practice to supplement prayers for safe deliverance by ringing church bells during thunderstorms, often with unfortunate results to those holding the bell ropes. One German author[1] states that in a thirty-three year period 386 churches were struck resulting in the deaths of 103 bell-ringers. The Campanile of St. Mark in Venice was destroyed or severely damaged nine times before a lightning conductor was installed in 1766. In one night in 1718 twenty-four church towers in Brittany were damaged by a single thunderstorm. These are but a few examples.

When artillery first came into widespread use gunpowder was often stored in the vaults of churches sometimes with disastrous consequences. At Brescia in 1769 100 tons of gunpowder were stored in the church of St. Nazaire when it was struck. The resulting explosion killed 3000 people and destroyed one sixth of the town. There are many cases such as this on record.

On the other hand many ancient buildings have never been struck by lightning. This is because some form of lightning conductor has been incorporated in the structure accidentally. Usually the roof has had a metal covering of some sort, connected to the ground by a metal rainwater pipe or staircase, or there has been some other such fortuitous circumstance.

In the days of wooden ships lightning was a great hazard. It appears that during the period 1799 to 1815 some 150 naval vessels were damaged.[1] Ten were completely disabled, one ship in eight was set on fire, and 200 men were killed or injured. The damage was estimated at £100,000. Several ships were lost with all hands during thunderstorms and one of these losses is known to have been due to a direct lightning stroke.

Nowadays with metal ships and the use of lightning arresters lightning damage is not nearly so widespread. However, it is still a hazard to aircraft which usually try to avoid thunderstorms. With metal aircraft the direct effect is not usually serious except to the radio and aerial if these are not grounded to the airframe. There is an associated danger, of course, due to the extreme turbulence which exists in thunderclouds and the possibility that under these difficult flying conditions the pilot might be temporarily blinded by the extreme brilliance of a flash.

Apart from occasional deaths of people caught in the open or sheltering under trees the most serious effect of lightning occurs, for example, in the vast timber forests of North America. During the hot dry summers the danger of fire is very great. Millions of acres are devastated by fires caused by lightning, resulting in many millions of pounds worth of damage every year.

3. The Lightning Conductor

The first serious study of thunderstorms was made by Benjamin Franklin in the middle of the eighteenth century. Franklin was a remarkable man. Born in Boston, Massachusetts, of English parents, he left home at the age of seventeen and became a highly successful printer and later diplomat. He was one of the signatories of the Declaration of Independence. Franklin had observed similarities between sparks obtained with his Wimshurst-like machine and lightning, and resolved to find out whether thunderclouds were charged. He therefore proposed that a long metal

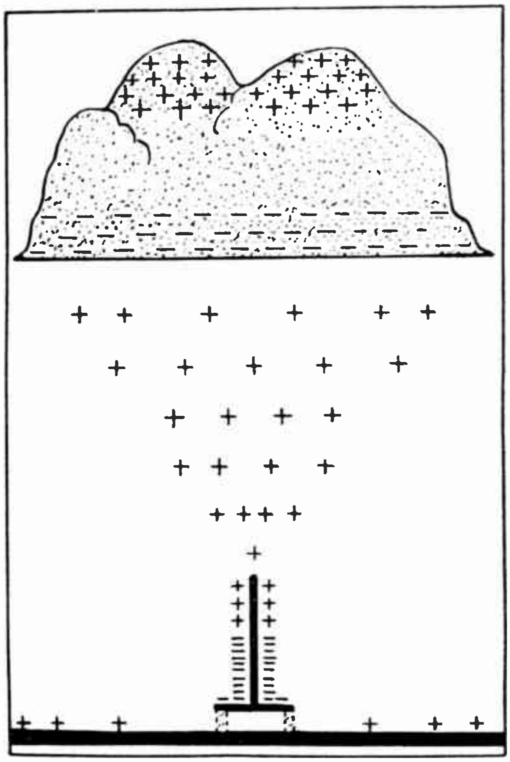

Fig. 1. Franklin's prediction of the charge distribution between thunderclouds, a long vertical metal rod isolated from the ground, and earth. (After Schonland.[1])

(Drawn by Dr. Mary Schonland)

Fig. 2. An impression of Franklin's experiment to establish his theory of charge distribution on thunderclouds. (After Schonland.[1])

rod, pointed at the upper end, should be mounted vertically on a high tower and insulated from the ground. If the cloud is, say, negatively charged Franklin predicted that the top of the rod would become positively charged and the bottom negatively charged as shown in Fig. 1. The positive charge would then stream off the sharp point leaving the rod with a considerable net negative charge. If an earthed conductor is then brought near the bottom end of the rod a stream of sparks should be obtained. The experiment was performed in France with the successful results illustrated in Fig. 2. Not having a high tower available Franklin himself obtained the same result with a kite. In further experiments he concluded, quite correctly, that (the base of) a thundercloud is usually negatively charged. These highly dangerous experiments were reported by many other people. There is a report where sparks ten feet long were obtained in one experiment, producing more noise than gunfire.[1] The experiments were also a source of amusement, particularly, it seems, to Louis XVI who ordered 200 monks to hold hands. They were given shocks, apparently "with prodigious effect". In other experiments several deaths occurred.

Franklin then proposed that, if connected directly to the ground as in Fig. 3, his rod might serve to protect buildings from lightning. He postulated that the upward discharge of electricity from the point would neutralize the charge on the cloud. We now realize of course that the amount of discharge from a single point can have negligible effect on the charge contained in a thundercloud perhaps extending for several miles. Nevertheless the lightning rod was found to be remarkably effective. In fact it simply conducts safely to ground a discharge which would otherwise pass through, and damage, a nearby structure.

Unfortunately a great scientific controversy now arose as to whether pointed or blunt conductors should be used. The advocates of blunt conductors maintained that points were dangerous and would attract a lightning stroke which would not otherwise occur. The quarrel grew very bitter and assumed political proportions as often happens in cases of this sort. At one stage George III was at loggerheads with the Royal Society. The situation was aggravated by the fact that many installations were inadequately grounded with the result that damage was caused to the buildings which were supposedly protected. A great variety of complicated contraptions were put on the market at this time and all were claimed to

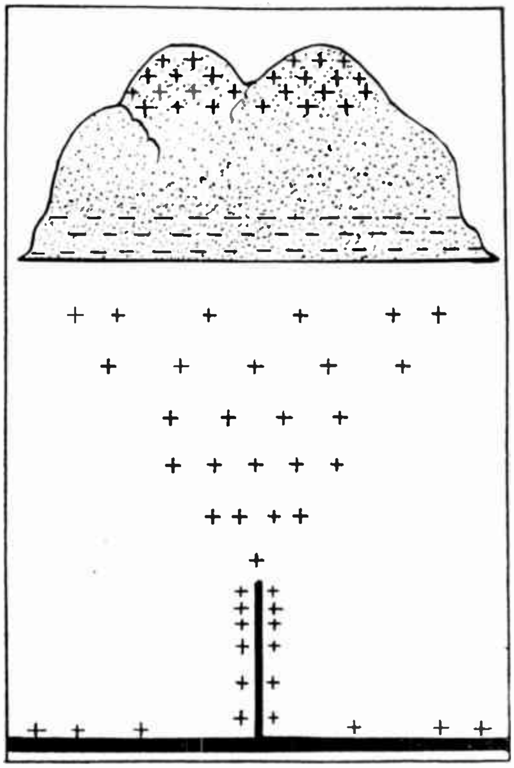

Fig. 3. Franklin's theory for the operation of a lightning conductor. (After Schonland.[1])

have miraculous protective powers. Even in those days it was doubtless essential to have a more elaborate installation on the roof than one's neighbour. One such is known to have been demolished by peasants who claimed it was the cause of the prevailing drought.[1]

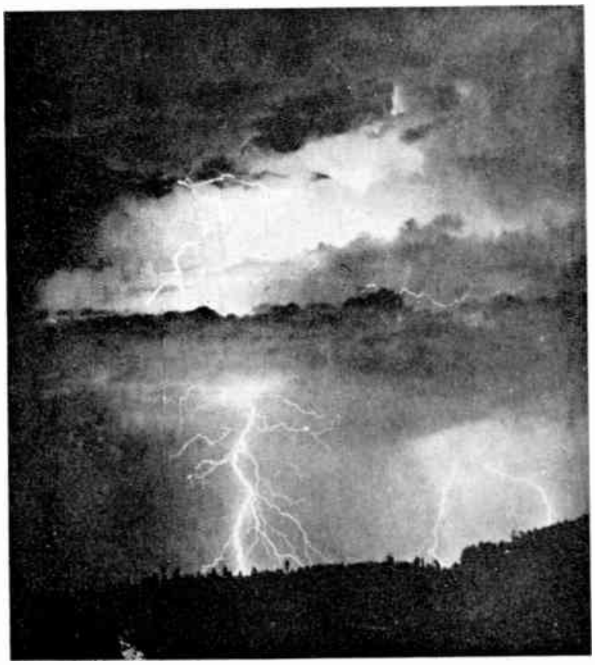

Fig. 4. Still photograph of lightning discharge to ground and within the cloud. (After Meek and Craggs.[4])

4. The Mechanism of the Lightning Stroke

Having dealt with some of the fancies, let us now consider some of the facts about lightning. About 44 000 thunderstorms occur somewhere in the world every day so that there are approximately 2000 occurring at this moment.[3, 4] From these storms there occur in the region of 100 strokes per second, corresponding roughly to a rate of energy dissipation of about 500 000 megawatts, i.e. more than three times the total world generating capacity! The amount of charge involved in a lightning stroke is about 10 coulombs and the peak current is in the region of 50 000 amps. A typical still photograph of a discharge to ground and within a cloud is shown in Fig. 4. The erratic nature of the path, and the multiple branching, can be clearly seen.

If there is a strong wind blowing, or if the camera is moved horizontally while the shutter is open, a time-resolved picture is obtained. It is then observed that each discharge is made up of several strokes which usually follow the same path. The entire discharge might last for up to 1 second but the individual strokes are visible for only a few milliseconds. The results obtained with a specially-constructed time-resolving, or "streak", camera are shown in Fig. 5 and illustrated diagrammatically in Fig. 6.

Figure 6 (*a*) represents a still photograph of a typical discharge. The idealized streak photograph at (*b*) shows that each stroke consists of at least two distinct processes. The discharge starts from the cloud at P and a "leader stroke" progresses towards the ground in a series of steps. Each step is about 50 m in length and there is a pause of about 40 μs between steps. The tip of this stepped leader is brightly luminous and the ionized channel behind it is faintly luminous. After each step the leader stroke changes direction slightly and it is this which causes the erratic nature of the path. When, after about 10 ms. the leader is within about 50 m of the ground the "return stroke", or "return streamer" jumps the gap and passes upwards to the cloud in a few tens of microseconds. Current persists in the very bright return streamer channel for perhaps 0·5 ms as the cloud discharges to ground. The channel diameter is about 6 inches and the temperature is probably in the region of 50 000° K.

After a time interval of about 40 ms another leader stroke appears but since the channel is partially ionized, the second and subsequent leader strokes progress continuously although they still have a brightly luminous tip. These are called "dart leaders." The leader strokes are always followed by a return stroke, which carries the charge from the cloud to the ground.

At (*c*) is shown a representation of the dart leaders obtained when the electric field strength is higher than usual.

The theory of the lightning discharge is rather involved and a detailed account would be out of place here. The observational difficulties are considerable, of course, since lightning occurs (apparently) randomly in both space and time and under conditions which are hardly conducive to accurate laboratory measurement. The "laboratory" in fact is the centre of a thunderstorm. The inventor of the first time-resolving camera, Sir Charles Boys, was particularly unfortunate and tried for 30 years without obtaining a suitable record. However, a general outline of two of the more recent theories may be given.

Schonland has proposed[5, 6] that when the electric field at a cloud reaches the breakdown strength of air ($\sim 3 \times 10^6$ V/m) a "pilot streamer" starts. In the case of a negatively-charged cloud base electrons start to flow from the cloud towards the ground. These produce ionization in the pilot streamer channel but there is a net negative space charge because of the charge flow from the cloud. This negative charge causes a reduction in the electric field at the cloud

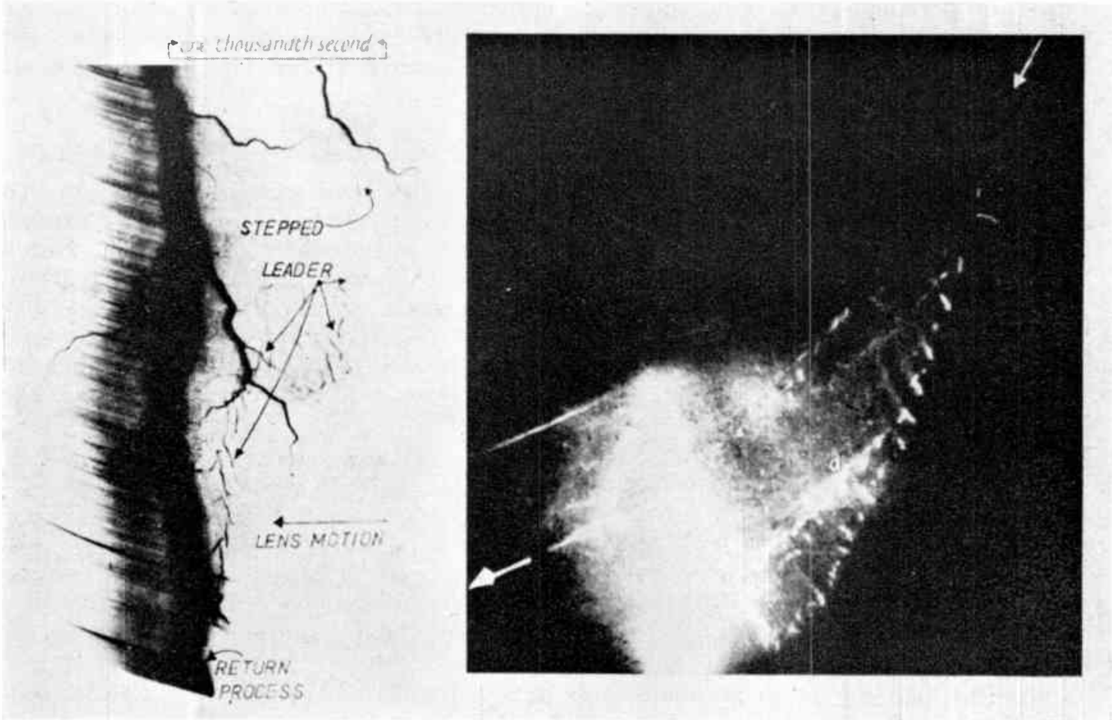

Fig. 5. Stepped leaders to first strokes. (After Schonland.[1])

end of the channel so that when the pilot streamer has advanced about 50 m current flow from the cloud ceases. With the disappearance of the highly mobile free electrons by capture and by their movement towards the front of the streamer a positive space charge, due to the relatively immobile positive ions, develops near the cloud. The resulting electric field very quickly becomes great enough to cause a further breakdown into the now partially-ionized pilot streamer channel. Charge thus flows suddenly into the pilot streamer right up to its front end and this short section develops into a highly-conducting arc-like channel.

There is now a high-conductance path extending from the cloud to the front of the original pilot streamer channel. Negative charge thus flows again towards the ground causing ionization forming a second section of pilot streamer. When the latter has progressed another 50 m or so the negative charge accumulated causes current flow to cease, the electrons are captured leaving a residual net positive charge so that breakdown occurs again into the

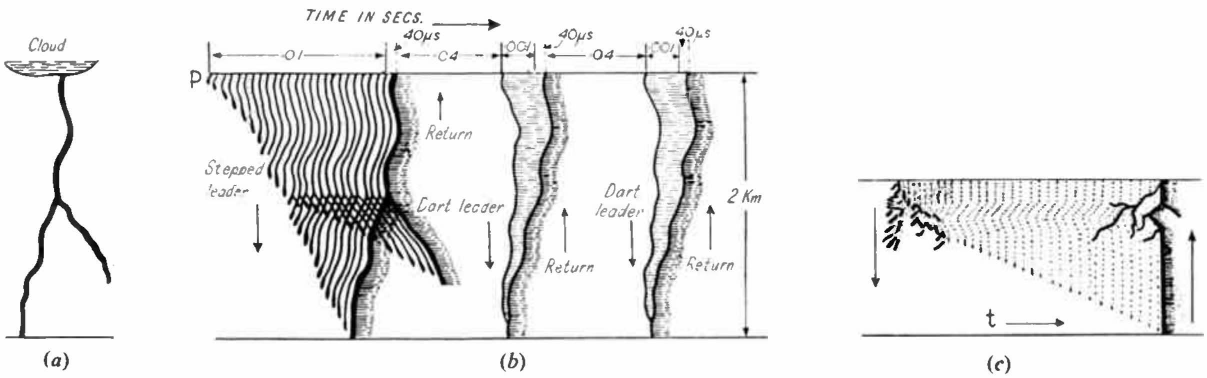

Fig. 6. (*a*) Fixed camera photograph of (*b*). (*b*) A three-stroke discharge to ground (with type α first-stroke leader) recorded on a camera with lens or film moving horizontally. (*c*) Type β first-stroke leader. (After Schonland.[3])

second section of partially-ionized pilot streamer channel. The arc-like section of channel, i.e. the stepped leader, increases in length very rapidly to the front of the second portion of pilot streamer and a third section begins to form. The stepped leader is thus preceded by a pilot streamer and travels to the ground in steps of about 50 m. The moderately-ionized pilot streamer is very faint in comparison with the bright, highly-ionized stepped leader so that it has not yet been photographed and the evidence for its existence is indirect.

When the stepped leader reaches the ground and links up with the positive leader, a conducting channel is formed between the cloud and the ground. A current wave then flows upwards and the cloud discharges to ground with the results which we all know so well. The pilot streamers are probably controlled in direction by local variations in the conditions existing in front of their tips, and their paths will be tortuous and branching, particularly when localized pockets of positive-ion space charge exist.

Another mechanism which has been suggested by Bruce[4, 6, 7] assumes that the diameter of the pilot leader is much less than the 2·5 m postulated by Schonland. The lateral voltage gradient to the surrounding air is thus much greater so that electron avalanches cause lateral corona currents to flow. The total current in the pilot leader thus increases with the length of the channel and it is proposed that when this current reaches a critical value a sudden transition to a thermally-ionized arc-type column occurs. In this way the leader stroke advances in a series of steps as before.

5. Radio Interference due to Lightning

One of the more spectacular effects of lightning upon radio reception occurs when a direct strike is received by an aerial which has not been disconnected from the receiver and grounded. Fortunately such an occurrence is rare and will not be discussed further! A much more common form of interference, known as precipitation static, often occurs as a preliminary to, rather than as a consequence of, lightning.

It is well known that if a pointed conductor acquires a sufficient charge, a stream of ions of the same sign as the charge flows from the point. In the intense fields which arise in a thunderstorm the resulting corona or glow discharge may be very conspicuous. It is thus quite common to see the tops of trees, buildings, bushes, ships, and particularly aircraft glowing brightly with this form of discharge which is known as St. Elmo's Fire. Apparitions of this kind, occurring as they do during violent thunderstorms, may well have caused the ancients some dismay. Such a discharge from an aircraft causes continuous interference that may blanket reception for long periods and this can be very serious of course. The effect is minimized

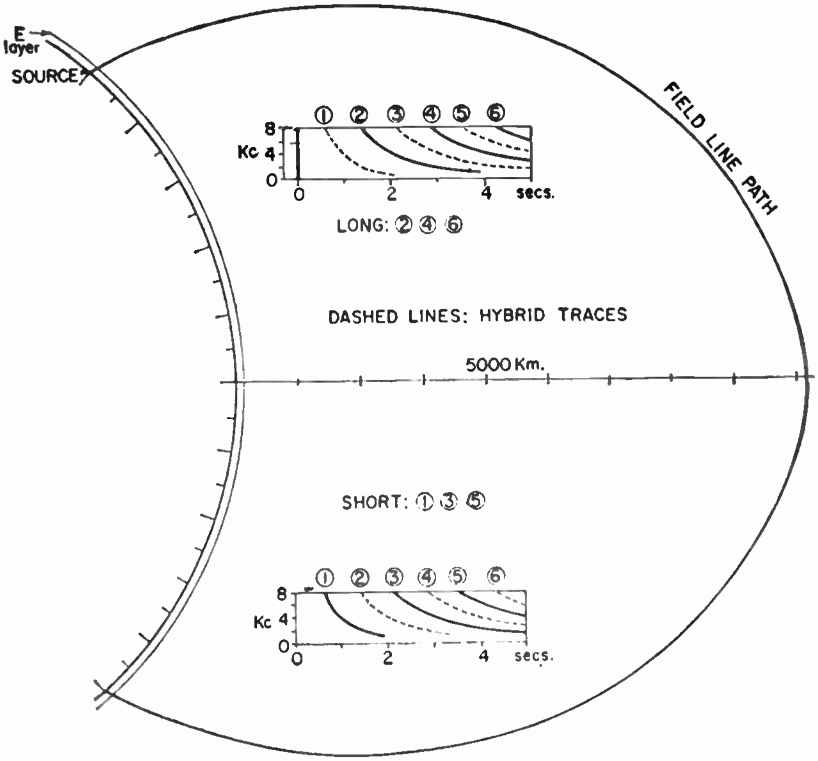

Fig. 7. From the lightning source at the Earth's surface radio waves enter the E layer of the ionosphere and become caught in the Earth's magnetic field, so that they travel far out into space along the field line path and return to Earth at the "conjugate point". There they are reflected along much the same path and may shuttle several times between the northern and southern hemispheres. The smaller diagrams show the changing notes of the whistlers and their arrival as recorded at receivers near the source and the conjugate point after alternate "hops". (This illustration and Fig. 8 are taken from "Whistlers", by R. A. Helliwell, *New Scientist*, **11**, p. 458, 24th August 1961.)

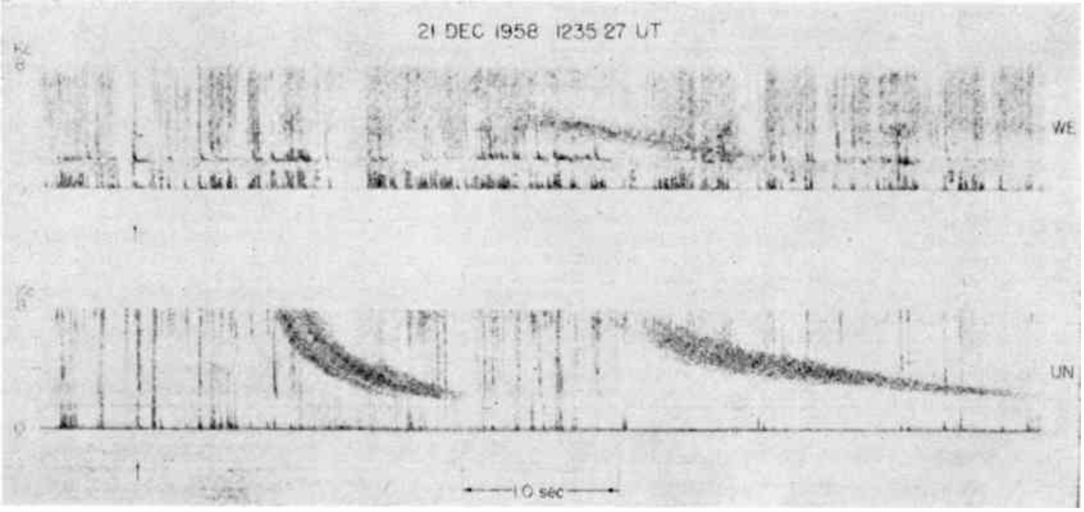

Fig. 8. Records of whistlers recorded at Wellington, New Zealand (near the lightning source) and at Unalaska, Aleutians (conjugate point). The intensity of different frequencies at different times is shown by the darkness of the display.

by providing points on the aircraft in positions where the interference caused is a minimum, sometimes by means of a trailing wire, and by using a shielded loop aerial since the induction field is mainly electrostatic.

The various strokes of a lightning discharge produce radiation. A nearby thunderstorm produces impulses of very high intensity in a radio receiver, while more distant storms give rise to a steady rattling or crackling. Such noise interference is usually called "static" or "atmospherics". By taking bearings on the source of the impulses and the static disturbances it is possible to plot the courses of storms, hurricanes, etc. The intensity of the static falls with increasing frequency, being negligible at u.h.f. and higher frequencies and is modified of course by the characteristics of the transmission path. The static level is, hence, usually higher at night when the propagation conditions are good.

At the lowest radio frequencies static is of very high intensity and for frequencies below about 30 kc/s sometimes takes a rather unusual form. If an aerial is connected to a high-gain audio amplifier strange whistling noises are often heard. These signals, called "whistlers", appear as a whistling tone of descending pitch lasting about 1 second.[8, 9] They may be repeated at regular intervals with the rate of fall in frequency becoming lower each time. Other v.l.f. emissions such as the "dawn chorus", "hiss", and "warblers" are often heard at the same time as whistlers but these latter noises are thought to originate in the ionosphere.

Whistlers, on the other hand, are produced by lightning, and their occurrence can be explained by reference to Fig. 7. When a discharge occurs at the point marked "source" some of the radiated energy may traverse the ionosphere where the waves interact with free electrons in the presence of the earth's magnetic field.[9] Propagation at very low frequencies can thus occur along the magnetic field lines which extend far out into space. When the propagated waves reach the other hemisphere some energy may reach the earth and some may be reflected. Dispersion of the signal occurs so that a range of frequencies up to, say, 8 kc/s emitted at $t = 0$ (Fig. 7), arrive at the conjugate point at different times, the higher frequencies travelling faster than the lower ones. A receiver at the conjugate point thus produces a whistling tone of descending pitch. This one-hop whistler is represented by the solid curve 1 in Fig. 7.

The reflected energy may re-traverse the field-line path and can be heard near the source as a two-hop whistler. Since the total path length has doubled, the dispersion is twice as great, as represented by the solid curve 2. Under certain conditions a long series of whistlers may be produced as shown by the solid curves of Fig. 7. In one instance a series of 25 reflections was still going strong when unfortunately the tape recorder stopped. The dotted curves represent components of "hybrid" whistlers which have been observed to be excited from opposite ends of the same path. A typical whistler recording is shown in Fig. 8 and was obtained during the International Geophysical Year.[9] The upper spectrogram was recorded at Wellington, New Zealand, and the lower one at the conjugate point of Unalaska in the Aleutians. The small arrows indicate the impulsive signal from the lightning discharge which travels to the receivers via the earth-ionosphere waveguide. The various components of the one-hop whistler at Unalaska are thought to have travelled along different paths.

A study of whistler propagation has yielded a great deal of information about the outer atmosphere and the ionosphere, and this research is being carried out with great vigour particularly in the U.S.A. Typical results so far show that (1) the electron density at five earth radii is about 100, (2) there is a large annual variation in the electron density in the outer atmosphere, and (3) the electron density in the outer atmosphere drops markedly after a large magnetic storm, and takes several days to recover. It has also been suggested that by transmitting high-power v.l.f. energy it is possible to modify considerably the electron distribution and energies in the outer atmosphere. Whistler mode propagation also suggests the possibility of communication with points in space without the limitation of a line-of-sight path.

6. Conclusion

In conclusion it may be stated that mankind's early terror of lightning has abated considerably although "Jupiter's thunderbolt" can never be contemplated altogether with equanimity. The mechanism of the lightning discharge makes a fascinating but difficult subject for study and much still remains to be discovered. In recent years lightning has ceased to be only a source of annoyance to the radio engineer and has provided knowledge of the earth's outer atmosphere which could not easily have been obtained by other means. Perhaps one day we may even learn how to harness some of the considerable energy associated with lightning and thunderclouds.

7. Acknowledgments

The author would like to make grateful acknowledgment to the excellent book "The Flight of Thunderbolts" (reference 1) from which much of the early part of this address was taken.

Grateful thanks are also due to the Director of the Radio Research Station for arranging the loan of the tape recordings of atmospherics and to Mr. Joyce, Senior Lecturer in Physics, Farnborough Technical College, for making available a portable van de Graaf generator.

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9. Demonstrations

The address was illustrated by the following demonstrations.

I. A metal sphere having a sharp point and representing a lightning arrestor was mounted on an insulator and shown to be uncharged by means of an electroscope. It was then brought near the high-voltage electrode of a van de Graaf generator which was meant to simulate a charged thundercloud. A distinct rustling sound was heard as charge flowed from the point. The generator was stopped and discharged. When the electroscope was now brought to the sphere a large deflection was observed showing that a charge is induced in an insulated (lightning) conductor by another nearby charged body (thundercloud), thus illustrating the principle of Franklin's experiment (see Figs. 1 and 2).

II. A small metal windmill having sharp points instead of sails was mounted on the high-voltage electrode. The "sails" rotated rapidly again demonstrating the discharge of electricity from charged points.

III. Long sparks were produced between the van de Graaf generator and an earthed electrode to illustrate the tortuous path and the erratic, impulsive nature of a long spark, which in this respect is similar to a lightning discharge.

IV. Tape recordings were played of various kinds of atmospherics such as whistlers, tweaks, dawn chorus, etc.

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The Practical Training of Professional Radio and Electronic Engineers†

A report prepared by the Training Panel of the Education and Training Committee of the Institution

1. Introduction

The Education and Training Report of the Education Committee[1] referred briefly to the subject of practical training. The present report is intended to be a more detailed guide to current practice in industry and to Institution policy with regard to practical training schemes for the professional radio and electronic engineer.

The definition of the professional radio and electronic engineer was included in the Education and Training Report and is repeated in Appendix 1 together with a summary of the main recommendations.

It is intended that separate reports should be concerned with the special aspects of practical training for technicians and craftsmen in the radio and electronics industry.

2. Type of Training Involved

Reference was made in the Education and Training Report to the three most important methods of obtaining academic education for radio and electronic engineers at professional level. These are:

(1) Full-time degree or diploma course.

(2) Sandwich training for the Diploma in Technology or Higher National Diploma.

(3) Part-time or block release training for Higher National Certificate and endorsement subjects.

The practical experience appropriate to these alternative means of academic education divide into two main types:

(a) (appropriate to methods (1) and (2) above)

A training scheme totalling two years which may be (i) entirely after graduation, (ii) divided into two one-year periods, as in the "thick" sandwich scheme (1-3-1), or (iii) divided into four periods, each of six months as in the "thin" sandwich scheme.

(b) (used in conjunction with method (3))

Five-year training period with part-time day or block release; this is now generally regarded as more suitable for the technician than for the professional engineer, but it still provides training for a number (albeit diminishing) of professional engineers.

Other possibilities exist, e.g. the practical training sandwiched between two periods of academic training, and are being tried out in one or two other subjects.‡ Their results will be watched with interest.

3. The Two-Year Scheme

During the last fifteen years the postgraduate training conducted by the radio and electronic industry has been under criticism for the rigid and unimaginative nature of the work. In the worst cases long periods were spent on routine processes in mechanical workshops. Normally, the university graduates and diploma holders are intelligent and enthusiastic, and the postgraduate training has to be arranged to provide continuous interest and intellectual stimulation and the exercise of some degree of initiative and responsibility. Some of the tuition will be given by technicians, but this should be limited to the demonstration of processes, and careful selection of the technicians will be necessary. The course should not be any longer than is absolutely essential to provide the breadth of experience desired.

In recent years the proportion of undergraduates on engineering courses taking such apprenticeships has probably fallen below 50%. There are, however, signs of improvement in this position since new and better schemes have been introduced.

It is now found by an ever-increasing number of employers that a course planned on a project basis[2] at all stages of the training is far more interesting and effective than the older types; the project idea is being introduced even for workshop training.

It is generally agreed amongst those consulted that every university graduate requires a period of practical training. This may sometimes be secured whilst in direct employment. Industrial employers seem to agree, however, that in most cases formal training is necessary and desirable; on the other hand, most Government Departments show much less enthusiasm for organized training, and none at all is provided for the Scientific Civil Service scientific officer class.

The actual length of training must depend upon the graduate himself, his previous experience, and his eventual employment. It is the Institution's view that a *basic* training scheme of about twelve months is ideal for the majority of graduates, and Appendix 2

† Approved by the Council for publication on 15th December, 1961. (Report No. 22.)

‡ e.g. Dip.Tech. course in Metallurgy, Battersea College of Technology, based on five-year course with practical training sandwiched between two periods in the College.

gives a break-down of a typical structure of the scheme. For some purposes, where wider experience is required, a two-year scheme may be necessary, but generally the second year training should be directed towards the graduate's eventual special field of employment.

4. "Thick" Sandwich Scheme

It is the considered opinion of many university teachers and most employers who have had experience in running a thick sandwich scheme that it has many advantages over the normal arrangement where a student enters university straight from school and follows up his university training with a two-year postgraduate apprenticeship. However, on the other hand, the period of a year's practical training before entering university is thought by some to be rather too long and a shorter preliminary training of six months might well be preferable; the argument is that the practical work should not get too far ahead of academic knowledge.

The advantage of practical training before entering university is most evident in the case of very young students; it helps to remove the "schoolboy" approach to study as well as giving a background of practical knowledge. An objection often raised is that it commits the young man to a particular firm or line of work too early, but the Institution sees no substance in this as students are not legally committed and most firms are commendably flexible in this matter.

5. "Thin" Sandwich Scheme

Although many employers seem to use a standard two-year postgraduate apprenticeship divided into six-monthly periods, it does not always provide the necessary correlation between academic and practical work at each stage. In the case of the Diploma in Technology scheme, it is the responsibility of the college authorities to ensure that the practical work carried out by the students is appropriate and sufficiently varied; it is also desirable that it should be related to the academic work as it progresses, and that lecturers make regular (even if infrequent) visits to their students in industry. It is probably necessary to develop the training so that it is based on projects which call on ingenuity and academic ideas for solution.

With the introduction of projects into the postgraduate scheme, it is sometimes possible to provide a common type of training for the postgraduate and sandwich type student.

It must be stressed that it is the co-ordination of practical and academic training which forms the appeal and potential strength of the thin sandwich course. It is therefore vital that every effort be made to ensure that this co-ordination is really effective. Much experiment will be necessary before it can be stated that success has been achieved.

Colleges commonly have a special mixed committee of industrialists and teaching staff to examine schemes of training from this point of view. It appears that most firms are very ready to meet the college criticisms and suggestions.

Proposals have been made in recent months for at least part of the practical training of a thin sandwich scheme to be taken in the college. A typical suggestion has been that all Dip.Tech. students should be college based for the first year with the first period of practical training being conducted in technical college workshops and laboratories. The second period of practical training would be arranged in industry when it is hoped that firms will adopt students as work based for the remainder of the course.

The advantage of college run practical training is in the close co-ordination of practical and academic work. The main disadvantage is the removal of the industrial environment.

It seems that some effort may be needed to keep a thin sandwich course as the basis for Diploma in Technology work. Administrative difficulties due to the uneven spread of the apprentice population over the year embarrass the employers, but most colleges are reluctant to adopt end-on courses.

The Institution strongly urges that these difficulties should be resolved since the basis of justification for the Diploma in Technology is the thin sandwich scheme and the close co-ordination of academic and practical training which it makes possible.

6. The Five-year Apprenticeship Scheme

It must be repeated that this general arrangement is designed for training technicians. However, a number of students training in this way have the ability to reach professional status and succeed in doing so. Many firms are prepared to give part-time day release for this purpose as long as the students make substantial progress in their studies.

As full-time and sandwich courses develop and become the accepted form of training for engineers (as also for high grade technicians), the number trained on a part-time day basis is bound to grow smaller and smaller. Already a large proportion of those students who reach the Ordinary National Certificate on a part-time day basis and secure distinctions in the O.N.C. examinations are routed on to Diploma in Technology and Higher National Diploma Courses, and some obtain scholarships to university.

Nevertheless the practical training secured on a five-year apprenticeship with part-time day release is extremely valuable, provided that it is well organized.

7. Vacation Training

For the university undergraduate and the full-time technical college student who is proposing to enter the radio and electronic engineering profession, vacation training should ideally be an essential part of the academic course, and all firms should co-operate in making suitable and well organized facilities available. The difficulties are appreciated, particularly those due to the return of sandwich students during this period. Five universities already include vacation training as an integral part of their degree courses and in many others recommendations are so strong that candidates rarely fail to undertake such training. There is no doubt that vacation training provides a very useful introduction to industrial employment, quite apart from the inherent value of the training itself.

Experience with vacation training shows that there is an enormous spread in the nature of the work and training provided, from first-class laboratory projects to routine maintenance and construction. This is not to say, of course, that all this training is not valuable. Only rarely is formal training provided, and it can be argued that it is really not only unnecessary but even undesirable, since one of the objects of vacation experience is for students to see industry *at work*. It is usually thought to be desirable for the first period of vacation training to be occupied in some heavy or mechanical industrial engineering to give the student some background, since electronic engineering is not typical of engineering generally.

It is very noticeable that bright students profit much more from this training than the duller ones, since they approach it with a great curiosity, and seek answers to all the questions which arise in their minds. Some organizations, indeed, are (perhaps because of this) very selective in their acceptance of students, taking only those assessed as being likely to achieve distinction in their studies. This is, however, very unfair on the colleges and universities and is deprecated. It has been found that only rarely is the student *not* looked after by an engineer of professional standing.

Vacation training also provides an opportunity for overseas experience, and such experience in seeing conditions in industry in other parts of the world is extremely valuable. More use could be made of the overseas vacation training schemes and the International Association for the Exchange of Students for Technical Experience already exists for this purpose.

8. The Problem of the Smaller Firm

Many of the schemes which have been outlined above employ the facilities of the larger firm with its education and training departments and apprentice supervisors. In radio and electronic engineering there are many smaller firms who wish to employ graduates but have no facilities of this type. There are several solutions to this problem, for example:

(a) *Co-operative schemes*

The combination of firms into a training unit, such as that organized by the Scottish Electrical Training Scheme Ltd.,[3] enables a number of small companies to give a breadth of experience to their apprentices which would otherwise not be available.

(b) *The use of technical college workshops*

To provide practical training on machine tools and in mechanical workshops it is normally essential to have special machines available. It is not always feasible to secure the right sort of training on machines which are engaged on full-time production. For this purpose many firms have apprentice training schools with a suitably equipped workshop. An alternative is to use technical college workshops and arrangements have been made between firms and local technical colleges for full-time or part-time practical training courses.

9. Other Types of Employment for which Formal Training is Desirable

The type of training which has been described is most essential for those radio and electronic engineers whose eventual employment will be in the manufacturing side of the industry, i.e. those concerned with the design, development and use of equipment for production. It is thus natural that most of these schemes are operated by the larger manufacturing firms in industry.

In the case of research workers, the need for formal training of this type is not so self-evident and it may well be that in many instances the time could be better used in research training, occasionally in the place of employment or more usually in the universities. But even so, some experience of design and production is useful to an engineering research worker and should be obtained whenever practicable. Certainly such experience is essential for those whose aim is a career in academic and technical teaching.

Electronic engineers whose eventual employment is in non-electronic industry provide a difficult problem. In many instances the industry concerned cannot give the electronic and general training which is desirable, and the employment of the electronic engineer is quite often in a "user" application only. It is to be expected therefore that the non-electronic firms wishing to employ electronic engineers would normally recruit engineers after they have had both their academic and practical training elsewhere, rather than attempt to provide the training themselves. Some of the very big organizations recruit a sufficiently large number of graduates and others for electronic work to justify a special training scheme.

Physicists are quite commonly employed in radio and electronic engineering and need thorough postgraduate training. They may also require additional academic training such as is provided at the various institutions and this can be made part of the postgraduate scheme.

10. Direct Employment

The Scientific Civil Service and many firms recruit graduates direct from the university into a position in a development or research laboratory. In such cases the graduate will often gain a great deal of valuable practical experience by contact with the various departments of the organization, especially if he is given a responsible task in which initiative is demanded and variety or breadth is involved. However, if variety of experience is provided on a regular basis for definite periods on each kind of work, then the training is likely to have a greater chance of success.

11. Specialized Postgraduate Courses

Many companies find it necessary not only to provide practical training for their graduates but also to give specialized postgraduate academic courses. These may be given by the firm's staff in the apprentice training school. In other instances special courses are organized by universities and colleges of advanced technology on a full time basis.

Special shorter courses in higher technology are run by many technical colleges and a considerable number of these are at genuine postgraduate level. It is quite often necessary to organize academic courses as part of graduate training, particularly where the university degree course did not deal in detail with radio and electronic engineering.

12. The Institution's Practical Training Requirements

The Institution recognizes that practical training is not common to all engineers and depends, as analysed above, on their eventual employment. For this reason no rigid practical *training* regulation has been written into the membership requirements, although adequate practical experience is required and this will, in many instances, include a formal practical training period. The majority of engineers would benefit by formal practical training and it is to be preferred, in most cases, to direct employment of the kind where the practical experience is secured largely on the initiative of the engineer himself. The most important exception is the research engineer, and in this connection the Institution would accept research training at a university, or in industry, as being equivalent to a formal postgraduate apprenticeship for the purpose of assessing a candidate's eligibility for membership.

It is not envisaged that applicants without practical training will be excluded from corporate membership of the Institution. However, it may be necessary in the future to distinguish, in terms of period of experience required, between those applicants who have had and those who have not had acceptable practical training.

13. Acknowledgments

The Education and Training Committee particularly acknowledges the assistance given by the following persons who have helped in providing information for this report:

Mr. L. P. Grice, Training Officer, General Electric Company, Coventry; Mr. P. C. Hordern, Secretary of the Appointments Board, University of Birmingham; Dr. J. W. Rowe, Technical Training Officer, Atomic Weapons Research Establishment; Mr. C. Stokes (*Member*), Principal Lecturer in Telecommunications, College of Advanced Technology, Birmingham; Mr. S. J. Whatling, Technical Training Officer, Kelvin & Hughes Ltd.

In addition, thanks are due to the many companies who completed questionnaires on their training schemes which provided the basis for the appendices, and all the past students who answered questions on their postgraduate training.

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"The relation between part-time, full-time and sandwich courses: 2. The point of view of the engineering industry", W. C. Cooper.

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16. Appendix 1:

The Recommendations of the Report on The Education and Training of the Professional Radio and Electronic Engineer[1]

(1) There should be a considerable expansion of education facilities for radio and electronic engineers, particularly in full-time and sandwich courses.

(2) There must be an increased emphasis on radio and electronics in electrical engineering and physics courses and there must be more new courses specifically designed for the radio and electronics profession.

(3) The Institution should continue to regard the Higher National Certificate with credits and suitable professional endorsements and appropriate radio and electronics content as a means to professional qualification.

(4) Practical experience under a qualified engineer is an essential part of the professional engineer's training but formally organized practical training is not the only means of achieving this.

(5) As far as practicable the training provided by the Services should be related to civilian courses and qualifications.

(6) Postgraduate courses in universities and technical colleges should be extended and industry should be encouraged to recognize the additional value of this training.

(7) Postgraduate research training in universities or colleges may be particularly valuable for any person who is to occupy an important place in industry.

This report also defined the professional radio and electronic engineer as follows:

"The professional radio and electronic engineer is competent by virtue of his education and training to apply scientific method to the analysis and solution of technological problems in his own field. He is capable of following progress in radio and electronic engineering by consulting and assimilating newly published information and by applying it independently. He should thus be able to make contributions on his own account to the advancement of technology. His work involves personal responsibility based on the exercise of original thought and trained judgement; he may be responsible for the supervision of technical and administrative work of others in the field of research, development, design and production or in the control of technicians engaged in the maintenance and operation of technical equipment."

17. Appendix 2:

Typical Formal Basic Postgraduate Training Scheme of Twelve Months' Duration

This is intended for graduates or diploma holders who have completed full-time courses and whose eventual employment is in the design or development laboratories of manufacturing electronics industry. It is assumed that during the academic training the candidate undertook vacation training of about twelve weeks duration.

Workshop training	2 months
Production and test	2 months
Installation and maintenance	2 months
Drawing office	1 month
Commercial department	1 month
Design and development	4 months

Where a graduate's eventual employment is not to be in design and development, other experience should be substituted in the last two months.

The nature of the experience under each heading will vary according to the types of organization and the equipment produced.

18. Appendix 3:

The Result of a Survey to Determine the Number of Radio and Electronic Engineers Employed and under Training

A questionnaire was circulated to just over two hundred firms. This questionnaire asked for information which was not easily obtainable. As a result only fifty-two returns were completed which could be fully taken into account in the analysis, whereas a further thirty were received which were either nil returns or incomplete.

Table 1

Summary of Returns

Size of firm	Number of returns	Total number of employees	Number of qualified professional engineers	Number of engineers of similar status but unqualified	Annual number of graduates recruited for practical training	Annual number of students recruited for thick sandwich training	Annual number of students recruited for thin sandwich training:		Annual number of students securing H.N.C.
							Dip.Tech.	H.N.D.	
Greater than 3000	12	90 000	1145	428	127	43	67	60	135
2000–3000	2	5000	81	76	5	8	2	6	7
1000–2000	5	7000	97	113	7	2	11	2	22
500–1000	10	6760	374	275	25	5	4	4	36
100–500	24	5727	160	214	5	5	11	4	37

The details obtained from firms employing less than one hundred persons contained so little information regarding training and the employment of professional engineers that this category has been omitted.

As far as the non-electronics industries were concerned, again very little information was secured. This was probably because of the interpretation of the term "radio and electronic engineer" in these industries.

On the sample of returns obtained and analysed, the contribution of the smaller firms is not significantly different from the bigger organizations on the basis of the number of employees and qualified staff. However, it is apparent that the larger organizations do devote more attention and effort to the training of their employees. In particular the larger firms provide a very much higher proportion of students on sandwich schemes.

It is also interesting to note that the larger firms have a smaller percentage of unqualified engineers, and the smaller firms are the biggest employers of engineers in professional positions who do not hold the normal qualifications.

The sample in some of these categories of firms is so small as to make the figures unrepresentative. The proportion of qualified engineers to total employees varies considerably with the type of organization and its product. This again determines the training policy and number of graduates recruited annually.

The above Report formed the basis of a Discussion Meeting on "Practical Training", sponsored by the Education and Training Committee and held at University College, London, on 27th September, 1961. Five formal contributions dealing with different aspects of the subject were presented at the meeting, namely:

"The Co-ordination of Academic and Practical Training"—E. MAY.

"Practical Training for Sandwich Diploma Courses"—B. F. GRAY.

"Vacation Training for University Electrical and Electronic Engineering Students"—J. C. CLULEY.

"Student Reaction to Formal Postgraduate Training"—A. J. KENWARD.

"Electronic Training in and for Non-Electronic Industry"—H. ARTHUR.

These contributions will be published in the April *Journal*.

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Tunnel Devices as Switching Elements

By

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AND

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Presented at a joint meeting with the Institute of Physics on "Tunnel Diodes" held in London on 7th February 1961.

Summary: The paper discusses the various ways in which the "gain" of a tunnel diode may be described and a straight line approximation to the tunnel diode's characteristic is used in obtaining an estimate of the switching time. An arrangement using ten diodes in series is suitable for decimal counting and circuits for a simple multiplier are given. Synchronous and asynchronous logic is compared and binary counting circuits are described. Two designs for a two-phase shift register are given. Mention is made of the part that backward diodes can play in tunnel diode circuits.

1. Introduction

In choosing a switching element to perform a given operation, the circuit designer is faced with a bewildering number of choices. He must choose the device that achieves a required standard of performance with the minimum of cost. Naturally, the circuit designer would like to know the position of the tunnel diode in relation to other circuit elements that are at his disposal. This position is not yet clear for two reasons:

(1) switching and logic circuit designs using tunnel diodes are not well established;

(2) the cost of tunnel diodes made by mass production methods is not yet known.

This paper is concerned in the main with the first of these problems. The answer to the second problem is, to a large extent, determined by the answer to the first problem because then the number of tunnel diodes in a given system may be found.

The tunnel diode differs from most other switching elements in one important respect—precision. It has at least one point on its characteristic (the current maximum, see Fig. 1) that can be defined with an accuracy better than 10%. Perhaps eventually with an accuracy of 1%, but this remains to be seen. Most other active circuit elements have their characteristics defined on a "greater than" or "less than" basis and tolerances on most parameters are often as much as 2 to 1. It is also probable that with a suitable control of the materials used in the tunnel diode, the voltage V_P and the voltage V_F, at which the current again equals the peak current, may be controlled with an accuracy of 10 to 20%. The circuit designer has a new degree of freedom in that he can choose the value of I_P and design circuits based on different values of I_P in a similar way that he can design circuits that depend on different resistor values. With transistors, on the other hand, it is usually impractical to design circuits that depend on different values of gain. An example illustrating this point is given in Section 3.2.

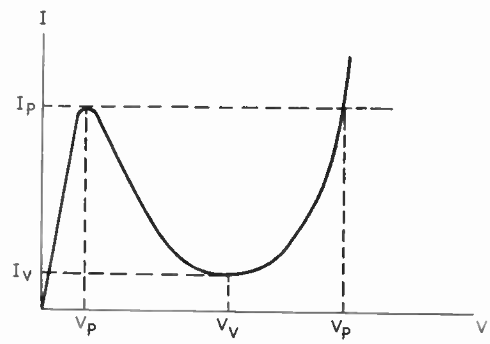


Fig. 1. Typical diode characteristic.

2. Characteristics of Tunnel Diodes as Switching Elements

2.1. *Energy Gain*

In many switching and logic circuits it is not sufficient for the circuit element to perform the operation; it should also be capable of providing an output that will operate a number of other similar circuits. Thus one must consider the amount of energy required to trigger the circuit in relation to the energy available from it when it has been triggered. A solution in the most general terms is not possible because the energy required to trigger the circuit of Fig. 3 is a function of the circuit tolerances. This is illustrated in Fig. 2 where I_P is assumed to have a 10% tolerance, R_L has a 5% tolerance and 5% is allowed as a margin. In the worst case the bias current will be 35% below I_P, at the other extreme it will be 5% below I_P.

† West Ham College of Technology; formerly with Standard Telephones and Cables Ltd.
‡ Standard Telephones and Cables Ltd., Transistor Division, Footscray, Sidcup, Kent.

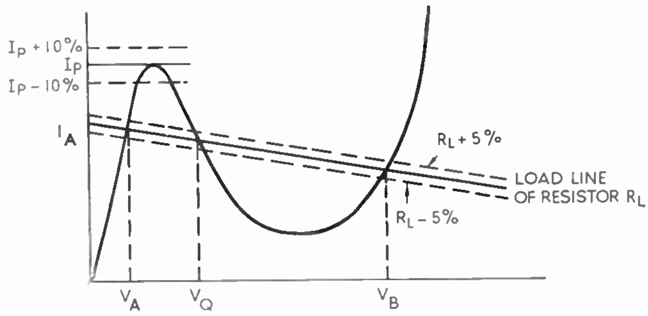


Fig. 2. Characteristics showing the effects of tolerances on the position of the load line.

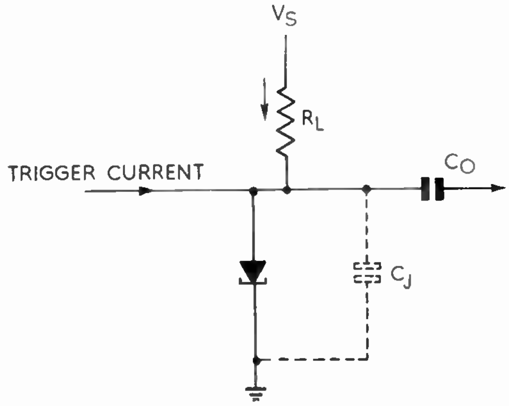


Fig. 3. Equivalent circuit when considering energy gain.

The energy required to trigger the circuit is

$$W_T > \tfrac{1}{2}C(V_Q - V_A)^2 + I_T \int_0^{t_T} \Delta V \mathrm{d}t \quad \text{......(1)}$$

where C is the junction capacitance of the diode, c_j (assumed invariant) plus any additional capacitance C_O across the diode.

V_A is the voltage at the bias point.

V_Q is the voltage at the point where the load line intersects the characteristics in the negative resistance region.

I_T is the amplitude of the constant-current trigger pulse.

t_T is the time taken for the voltage to increase from V_A to V_Q.

ΔV is the change in voltage across the tunnel diode at any time, t.

Q is the point of "no return", i.e. if the trigger current is removed before the voltage across the diode reaches V_Q, the diode will (in the absence of stray inductance) return to its original state.

The second term in (1) is not negligible unless the trigger current is very large compared with I_P (see Sect. 2.4—switching times) and a general expression for the energy gain is not easy to obtain. However, if one neglects the second term in eqn. (1) it is possible to obtain an expression for the maximum possible energy gain.

It is assumed that the output energy is used to charge a capacitance C_O where $C_O = C - c_j$ and c_j is the diode junction capacitance (see Fig. 3). The output energy is

$$W_O = \tfrac{1}{2}C_O(V_B - V_A)^2$$

and the energy gain is

$$W_g = \frac{W_O}{W_T} < \frac{C_O}{C}\frac{(V_B - V_A)^2}{(V_Q - V_A)^2}$$

This will be a maximum for C_O large compared with c_j. (However a large value of C_O will result in a long switching time.) It will be greatest for the largest value of V_B, therefore a material with a large energy gap is desirable. It is of interest to note that the energy gain is not, to the first order, dependent on the peak current.

The value of W_g for a germanium tunnel diode, assuming $V_P - V_A$ = 60 mV, $V_B - V_P$ = 450 mV and $C_O > c_j$ is 56 times. This result is somewhat artificial in assuming a constant input current and a purely capacitive load.

2.2. *Power Gain*

One can also consider the power gain under purely static operating conditions. In Fig. 4, to make the circuit switch from A to B, the load line representing the resistance R_L must be shifted from the position shown by the firm line to the position given by the dotted line. To do this, the increment in input power is

$$P_I \simeq R_L(I_P - I_A)^2 + (V_P - V_A)(I_P - I_A)$$

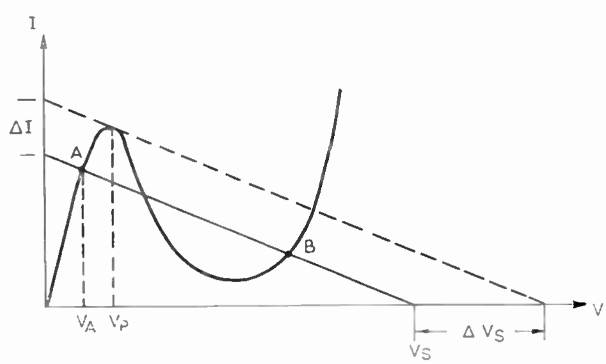


Fig. 4. Showing how the load line is shifted to enable the power gain to be calculated.

When the circuit switches, the power in the load changes from

$$(V_S - V_A)I_A$$

to $$(V_S - V_B)I_B$$

The power gain is

$$P_g = \frac{(V_S - V_A)I_A - (V_S - V_B)I_B}{R_L(I_P - I_A)^2 + (V_P - V_A)(I_P - I_A)}$$

Note that the minimum value of R_L gives the maximum gain. The minimum value of R_L is approximately

$$R_L \simeq \frac{V_V - V_P}{I_P - I_V}$$

A typical value of P_g for a germanium tunnel diode, assuming the tolerances given earlier, would be about 2 times for the most adverse tolerance and about 100 times for the most favourable tolerance. Power gain is more sensitive to circuit tolerances than is energy gain.

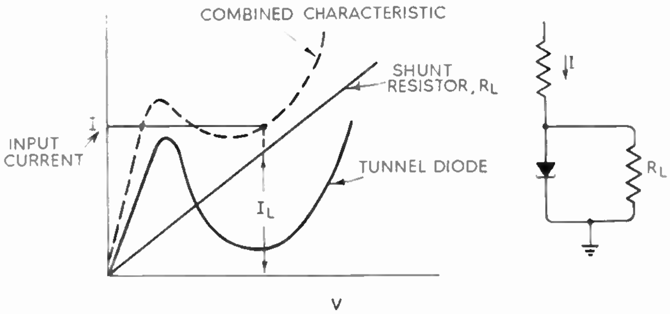


Fig. 5. Combined characteristic of a tunnel diode shunted by a fixed resistor.

2.3. *Current Gain*

By considering the characteristic of a tunnel diode shunted by a resistor (Fig. 5) it is possible to see that with a constant current supply the current in the load can never exceed a figure of the order $(I_P - I_V)$. In fact if tolerances are taken into account, the figure is likely to be of the order $\frac{1}{2}(I_P - I_V)$. This results in current gains of the order 1 to 2 times.

2.4. *Switching Times*

The time taken to switch from one side of the negative resistance region to the other is related to the product of negative resistance and shunt capacitance. The expression for a figure of merit is $t = C/g$ (where C is the total shunt capacitance (junction plus stray) and g is the value of negative conductance at an arbitrary point in the negative resistance region). This represents a figure that can only be achieved under ideal conditions and actual switching times may be much longer.

Figure (6) shows an idealized tunnel diode characteristic consisting of four regions. In region I the conductance is high and positive; in region II it is zero; in region III it is high and negative; in region IV it is zero, and in region V it is high and positive.

It is assumed that a trigger pulse is applied to the circuit which would under equilibrium conditions shift the static load line parallel to itself so that the difference between I_P and the point on the load line vertically above it is I_O. It is assumed that the duration of the trigger pulse is longer than the tunnel diode switching transient, i.e. it can be regarded as a step function of current applied to the circuit. It is also assumed that the diode is biased from a constant current source.

It may be readily shown that the total switching time is made up of five transients (see Fig. 6 for the meaning of the symbols)

Region I $\quad t_1 = \dfrac{C}{g_1}\log_e\left(\dfrac{I_T}{I_O}\right)$

Region II $\quad t_2 = \dfrac{C(V_{P2} - V_{P1})}{I_O}$

Region III $\quad t_3 = \dfrac{C}{-g_3}\log_e\left(\dfrac{I_O}{I_K}\right)$

Region IV $\quad t_4 = \dfrac{(V_{V2} - V_{V1})C}{(I_K)}$

Region V $\quad t_5 = \dfrac{C}{g_5}\log_e\left(\dfrac{I_K}{I_T}\right)$

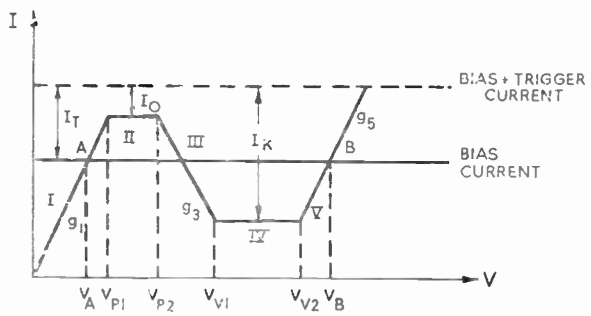


Fig. 6. Idealized characteristic to facilitate the calculation of switching times.

It is apparent that the value of t_2 is very dependent on the value of I_O. Hence the amplitude of the trigger pulse will have a marked effect on the delay time. The values of t_1, t_2, t_3, t_4 and t_5 are given below for a tunnel diode having the following parameter values:

$V_{P1} = 45$ mV; $V_{P2} = 55$ mV; $V_A = 40$ mV;
$V_{V1} = 250$ mV; $V_{V2} = 300$ mV; $V_B = 350$ mV;
$I_P = 5$ mA; $I_A = 4$ mA; $I_V = 1$ mA;
$I_T = 1{\cdot}5$ mA; $I_K = 4{\cdot}5$ mA; $I_O = 0{\cdot}5$ mA;

$$g_1 = 0{\cdot}1 \text{ mho}; \quad g_3 = -0{\cdot}02 \text{ mho};$$
$$g_5 = 0{\cdot}04 \text{ mho}; \quad C = 30 \text{ pF}.$$

Then $t_1 = 0{\cdot}31$ nanoseconds
$t_2 = 0{\cdot}60$ nanoseconds
$t_3 = 3{\cdot}3$ nanoseconds
$t_4 = 0{\cdot}33$ nanoseconds
$t_5 = 0{\cdot}83$ nanoseconds

$$t_1+t_2+t_3+t_4+t_5 = 5{\cdot}4 \text{ nanoseconds} = 3{\cdot}6\frac{C}{g_3}$$

If instead of the mean value of negative conductance, the maximum value is taken as $g_m = 0{\cdot}05$ mho, the actual switching time is nine times the figure of merit C/g_m. The computed figure for switching time seems to be somewhat larger than the figure obtained from measurements on actual diodes.

2.5. *Flexibility—Comparison with other Devices*

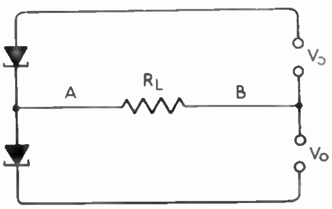
The above analysis has shown that there may be considerable problems in interconnecting tunnel diode circuits. The tunnel diode seems to be capable of giving a large energy gain, and works best when it is supplied from a current source and works into a capacitive load. A tunnel diode is not capable of providing a true current source to drive another tunnel diode. This points strongly to the use of *tunnel diodes with transistors*. In fact some of the most successful switching circuits built so far do contain transistors and tunnel diodes. The transistor is an ideal complement to the tunnel diode because, in the common emitter configuration, it behaves as a constant current generator and its input impedance has a capacitive reactance.

With the somewhat idealized assumptions made above, no account has been taken of losses associated with gating or routing networks, nor of the difficulties of obtaining a power match if one is needed. So in this sense the results are somewhat optimistic.

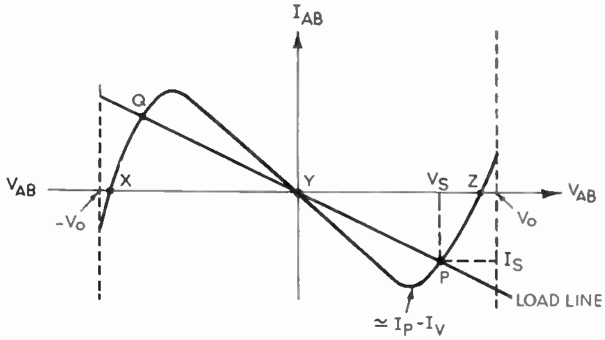
The conclusion is that, compared with the transistor, the tunnel diode is lacking in flexibility. Nevertheless the tunnel diode has some important applications; these will be considered in the next sections.

Compared with a *p–n–p–n* switching diode, the tunnel diode has greater uniformity of characteristics, a much lower power consumption, and is generally much faster. On the other hand, it has a current rather than a voltage maximum, and this is not always as convenient, especially in matrix store which is voltage- rather than current-operated.

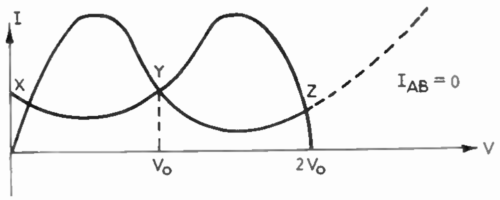
A *p–n–p–n* triode or a thyristor has the obvious advantage of a third terminal giving current gain; in general this makes for much greater flexibility. However, as the tunnel diode is the simpler device, it should be cheaper.



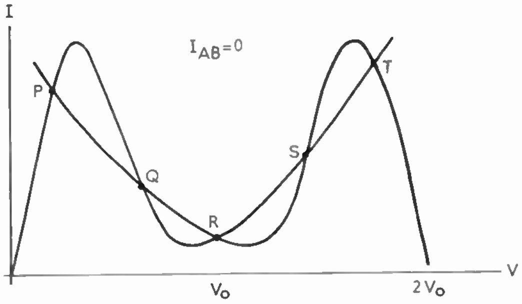
(*a*) Circuit.



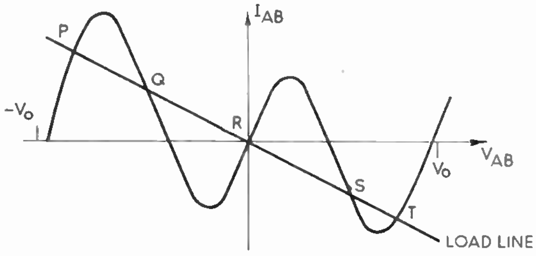
(*b*) Characteristic of a matched pair of diodes.



(*c*) Characteristics of a pair of diodes—one drawn as the load line for the other.



(*d*) Similar to Fig. 7 (*c*) but a higher voltage giving three stable operating points.



(*e*) Composite characteristic corresponding to Fig. 7 (*d*).

Fig. 7.

3. Tunnel Diodes in Pairs and in Long Chains

3.1. *Pairs*

There are a number of ways in which the characteristics of a pair of tunnel diodes may be represented graphically. For example, when two tunnel diodes are connected directly in series (Fig. 7 (*a*) with R_L infinite) one tunnel diode characteristic may be drawn as a load line on the other (Fig. 7 (*c*)), and there will be two possible stable states, X or Z. Referring back to Fig. 7 (*a*), the voltage current characteristic of the pair of diodes as measured at A (with B as reference) may be drawn as in Fig. 7 (*b*) with the centre point resistor as load line. Figure 7 (*b*) shows that, if the pair of tunnel diodes is well matched, the trigger sensitivity will be equal when triggering from P to Q or from Q to P. However, the sensitivity will be no greater than that of a single tunnel diode. The output voltage will be about the same as that of a single tunnel diode.

By increasing V_O so that the characteristics intersect (when one is drawn as a load line on the other) in five points, we obtain three stable points as in Fig. 7 (*d*).

Figure 7 (*e*) is the composite characteristic corresponding to Fig. 7 (*d*). It may be easier to obtain the type of behaviour if resistance is added in series with the diodes.

Tunnel diode pairs in circuits using synchronous or clock pulse operation are of great interest. The clock pulse is applied in such a way as to bias the diode pair as shown in Fig. 7 (*c*). Whether the system will operate at point X or point Z is determined by the relative values of the peak currents of the two diodes. The point of operation for a particular pair will always be the same one until a current is fed to the junction of the diodes. If the diodes are well-matched, the deciding current can be made very small. This system is referred to again in Section 4.1.2.

3.2. *Long Chains*

When a number of tunnel diodes are connected in series, as in Fig. 8, it is possible for successive input pulses to "switch on" one tunnel diode at a time.

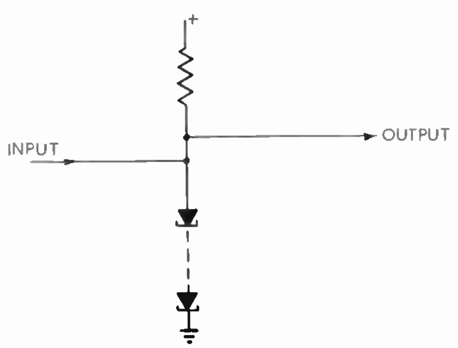


Fig. 8. Tunnel diode chain.

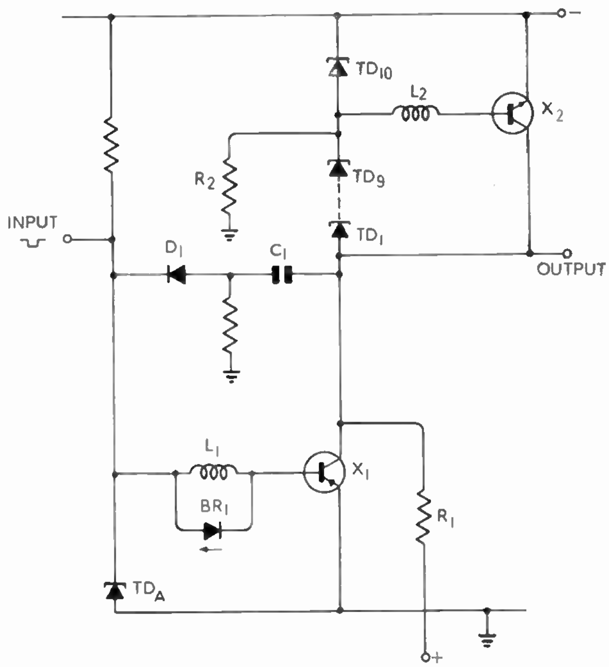


Fig. 9. Decade scaling circuit using a chain of ten diodes with unequal peak currents.

By "switch on" is meant that it changes from the low to the high-voltage state. There are two modes in which a circuit of this kind can work: it can be current-operated or can work on an energy-sharing basis.

In the current-operated circuit, the tunnel diodes are chosen to have unequal peak currents. We will assume for the purpose of this example that there are ten diodes having peak currents at 1 mA intervals from 1 to 10 mA. The standing bias on the circuit is less than 1 mA, and the valley currents of all tunnel diodes are less than the standing bias. (This last condition is essential in the particular circuit described, but there are other circuits where it would not be necessary.) Initially all diodes are in the low-voltage state. The input is in the form of a saw-tooth of current which increases in amplitude until the 1 mA diode switches, whereupon it is reset to zero. The next input will increase to 2 mA before it is reset, and so on. The tenth input will cause the whole chain to be reset to zero.

Figure 9 shows the complete circuit diagram of a decade scaler which embodies the above-mentioned principle.

The input pulse switches TD_A to the high voltage state. This step of voltage drives an exponentially rising current into the base of transistor X1 due to the action of inductor L1. This current is amplified by X1, and is transmitted to the tunnel diode chain.

As a diode in this chain changes state, it applies a positive pulse through C1 and D1 to TD_A, switching it back to the low voltage state. The backward diode BR1 by-passes L1 to provide a fast resetting action. The resistance R1 maintains the current in the tunnel diode chain. The tunnel diode TD10 has the highest value of peak current and is, hence, the last one to change state. On changing state, it applies a pulse to transistor X2, which in turn resets the entire chain. R2 provides supplementary bias to TD10 ensuring that, during the resetting of the chain, TD10 is the last diode to switch off. A small inductance L2 delays the chain resetting until no more current flows in X1. The chain reset pulse provides the output. D1 prevents this pulse from reaching TD_A. The ultimate speed of the circuit is determined by the speed at which the resetting operation can be carried out.

In the above example, it was assumed that the diodes had peak currents at 1 mA intervals from 1 mA to 10 mA. It should be noted that the tolerance requirements on the 10 mA diode are very much more severe at 2% than they are on the 1 mA diode, where 20% is acceptable. Of course, the range does not have to be 1 to 10 mA; it could be 5 to 10 mA at 0·5 mA intervals or 5 to 15 mA at 1 mA intervals. The choice of the best scale is determined by tolerances and peak-to-valley ratios. The intervals could have a logarithmic rather than a linear spacing.

In the other mode of operation† there is no attempt to select the tunnel diodes for peak currents. The input is a voltage pulse applied through a capacitor to the input terminal. For a certain range of input levels only one tunnel diode will switch at a time. This input is not as critical as may appear at first sight. The operation of this type of circuit is not well understood, but it seems likely (bearing in mind the calculations in Section 2.1) that the diode with the lowest capacitance will switch first. Once switching action has started on one diode, it will inhibit switching action on the other diodes, as there is only a fixed voltage available to be shared among the diodes in the chain. There is correlation between capacitance and peak current, but it is not perfect; and if selection is necessary, it will probably have to be on the basis of capacitance. There is a further difficulty with this circuit because, as the number of diodes in the high voltage state increases, the series resistance of the chain increases, and the circuit demands a larger trigger pulse.

The constant-current mode of operation has the advantage of being much more "designable". The energy-sharing mode has the advantage of being simpler and possibly faster but the number of diodes in a chain is limited. In very fast counting circuits it may be possible to use the energy sharing circuits for the input stage followed by constant current circuits for lower speeds.

† P. Spiegel, "High speed scalers using tunnel diodes", *Rev. Sci. Instrum.*, **31**, p. 754, July 1960.

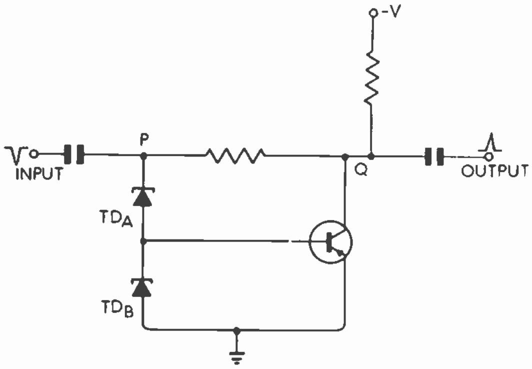


Fig. 10. Tunnel diode binary counting circuit.

Figure 10 shows a binary counting circuit. Each input pulse switches on only one tunnel diode. TD_A must switch on first. This can be ensured by letting TD_B have the higher capacitance. When TD_B is switched on, the transistor is turned on and switches off both diodes.

Hence TD_B must be the last diode to switch off. This will happen if its valley current is lower than that of TD_A. The output consists of a short positive pulse, requiring that the following stage contains an *n–p–n* transistor. If the circuit is required to operate a gate, it may be possible to take an output from point P rather than point Q. It may also be convenient to make TD_B a gallium arsenide tunnel diode which is better suited to turning on the transistor.

4. Logic Circuits

4.1. *Tolerance Problems in Threshold Logic*

The tunnel diode, being essentially a threshold device, is suited to logic gating. It can perform basic AND, OR and inversion functions.‡ The functional limit of the tunnel diode as a logic gate is provided by the tolerance on the diode characteristic as well as on the biasing components.

To illustrate this problem, calculations of input and output capabilities are shown in Appendices 1 and 2. Appendix 1 deals with an idealized static system of logic using one diode per gate, whereas Appendix 2 deals with a synchronous system of logic utilizing matched diode pairs.

‡ G. W. Neff, S. A. Butler and D. L. Critchlow, "Esaki diode logic circuits", *Trans. Inst. Radio Engrs (Electronic Computers)*, **EC-9**, No. 4, p. 423, December 1960.

R. H. Bergmann, "Tunnel diode logic circuits", *Trans. Inst. Radio Engrs (Electronic Computers)*, **EC-9**, No. 4, p. 430, December 1960.

4.1.1. Static logic

An idealized system of logic is shown in Figs. 11 (*a*) and (*b*). This is included only to illustrate the effect of tolerances. Both the inputs and the outputs take the form of a constant current of magnitude I on which there is a tolerance of $\pm\alpha$. It must be noted that in this system the tunnel diode is not biased and that it acts purely as a threshold gate which presents a low shunt resistance when the tunnel diode is in its low voltage stage.

n inputs should overcome the peak current of the diode which then acts as a high resistance and allows m outputs to be fed out of the system before the diode current falls below the valley current value, switching the gate off again. It is assumed that the input currents persist when the output currents are required.

The tolerance on the peak current of the diode, β, as well as α, plays a part in determining a minimum value of I which is given in eqn. (5) (Appendix 1) as

$$I_{\min} = 2I_P\left(\frac{\alpha+\beta}{1-\alpha^2}\right)$$

where I_P is the nominal peak current of the diode. The maximum number of inputs is then given by eqn. (6).

$$n_{\max} = \frac{(1+\alpha)(1+\beta)}{2(\alpha+\beta)}$$

The effect of tolerance on the number of input pulses is shown in Fig. 11 (*c*).

The number of outputs that are available is given by eqn. (7)

$$m = n_{\max} - 1 - \frac{(1+\beta)}{2A_{\min}(\alpha+\beta)}$$

where $A_{\min}$ is the worst peak-to-valley current ratio which can be expected.

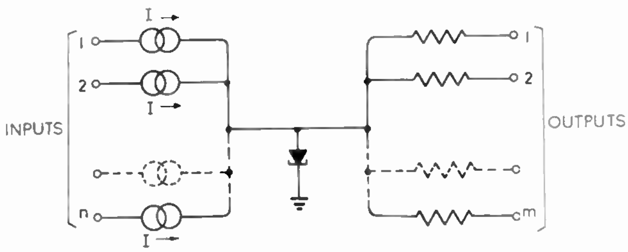
In operation this system would not be practicable, since the example assumes that the input circuit has a high impedance and the load has an impedance which lies between the gate resistance limits. This could not be achieved if the input and load circuits were tunnel diodes. In such a case, the situation with regard to fan-out and fan-in ratios is considerably worsened.

4.1.2. Synchronous logic

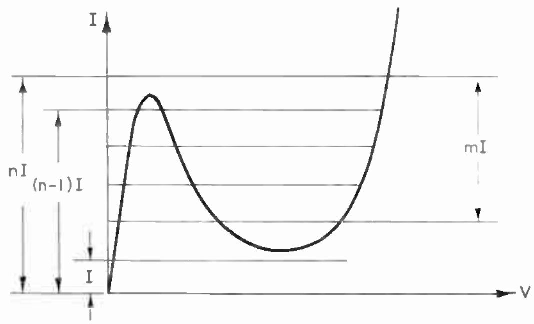
This system of logic uses the principles outlined in Section 3.1. The operating conditions are shown in Fig. 11 (*d*).

It is shown that n inputs are required to overcome the difference between the peak currents of the two diodes.

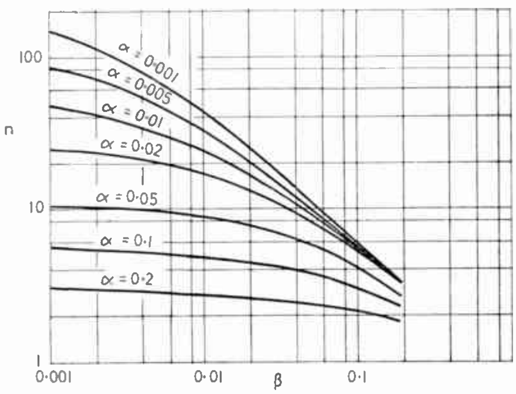
The expression for the maximum number of inputs



(*a*) Idealized static AND gate.

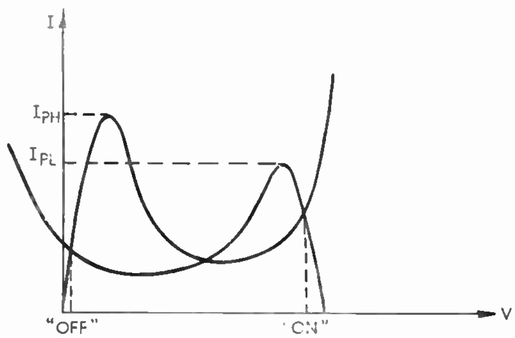


(*b*) Loading of static AND gate.



(*c*) Variation of n with tolerances.

Fig. 11.



(*d*) Synchronous AND gate operating conditions.

as a function of tolerances is given by eqn. (9) in Appendix 2:

$$n_{\max} = \frac{(1+\alpha)}{2}\left[\frac{(I_{PH}(1+\beta) - I_{PL}(1-\beta)}{(I_{PH}(\alpha+\beta) - I_{PL}(\alpha-\beta)}\right]$$

where $n_{\max}$ is the maximum number of inputs

α is the tolerance on the trigger current

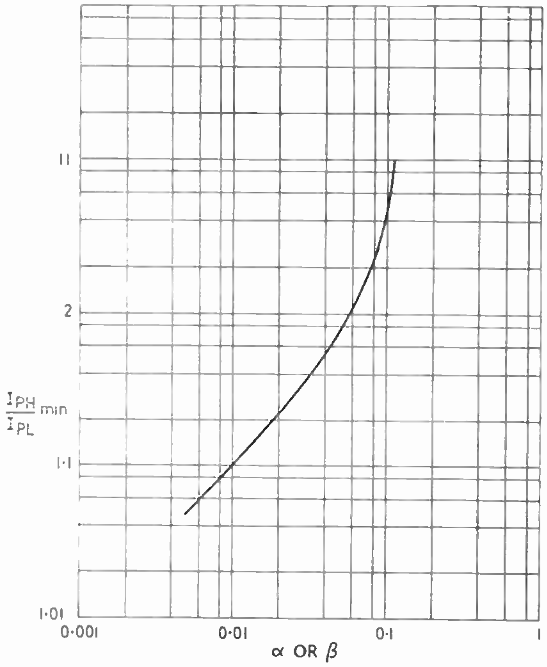
β is the tolerance on the peak currents

I_{PH} is the higher peak current

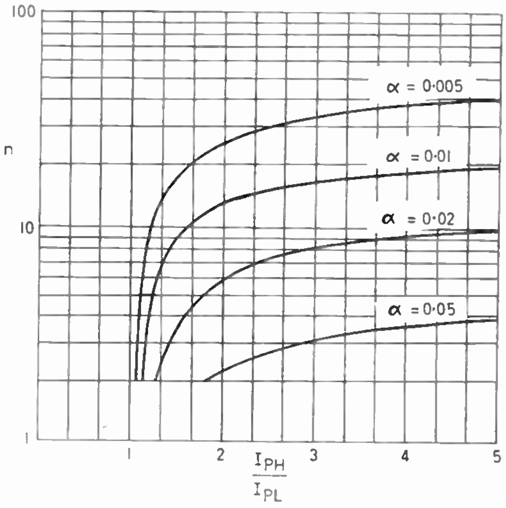
I_{PL} is the lower peak current.

The number of inputs is dependent on the ratio of the two peak currents. The larger the ratio, the more inputs can be applied. There is a minimum value that the ratio I_{PH}/I_{PL} can have. This occurs when $n = 2$ and is a function of the circuit tolerances. To illustrate this, assuming that $\alpha = \beta$, the variation of $(I_{PH}/I_{PL})_{\min}$ is shown in Fig. 11 (*e*). Figure 11 (*f*) shows the variation of the number of inputs with I_{PH}/I_{PL} for various values of circuit tolerance, again assuming that $\alpha = \beta$.

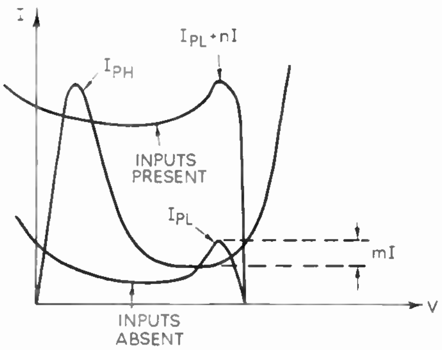
In a synchronous system of logic, it must generally be assumed that the inputs are de-energized when the gate supplies information to its output. Hence the number of outputs available (m) is dependent on the difference between the valley current of the diode with the higher peak current and the lower value of peak current. The latter must be greater in magnitude than the former. This situation is shown in Fig. 11 (*g*).



(*e*) Variation of $\frac{I_{PH}}{I_{PL}}$ min with tolerances.

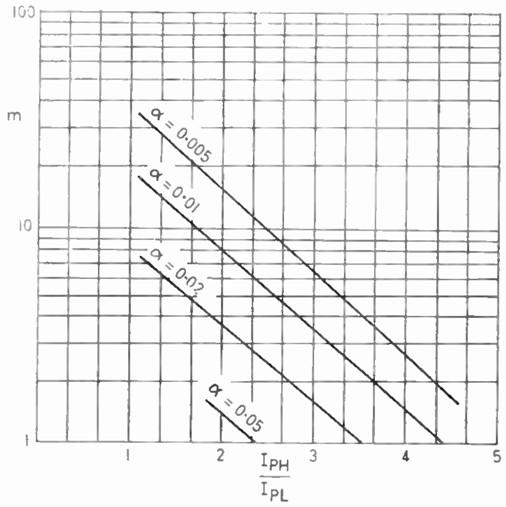


(*f*) Variation of n with $\frac{I_{PH}}{I_{PL}}$ and tolerances.



(*g*) Loading situation after the removal of inputs.

Fig. 11.



(*h*) Variation of m with $\frac{I_{PH}}{I_{PL}}$ and tolerances.

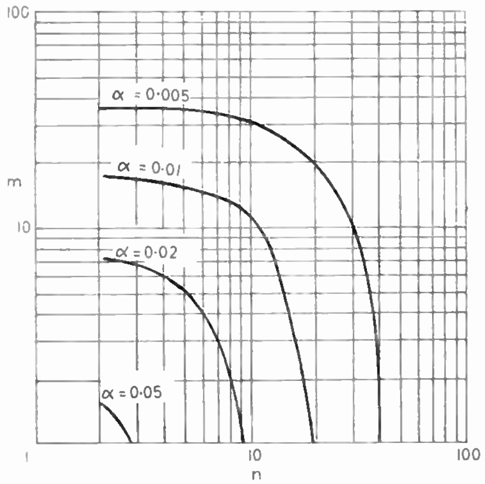


Fig. 11. (j) Variation of m with n.

The number of outputs is

$$m = \frac{(1-\alpha)}{2}\left[\frac{I_{PL}(1-\beta) - \dfrac{I_{PH}}{A_{\min}}(1+\beta)}{I_{PH}(\alpha+\beta) - I_{PL}(\alpha-\beta)}\right] \quad ......(2)$$

Assuming that $\alpha = \beta$, the variation of m with I_{PH}/I_{PL} is shown for various values of circuit tolerances (Fig. 11 (h)). It is evident that a compromise has to be made between inputs and outputs as, to increase the one, the other must be sacrificed.

The ratio m/n is shown to be

$$\frac{m}{n} = \left(\frac{1-\alpha}{1+\alpha}\right)\left[\frac{I_{PL}(1-\beta) - \dfrac{I_{PH}}{A_{\min}}(1+\beta)}{I_{PH}(1+\beta) - I_{PL}(1-\beta)}\right] \quad ......(3)$$

The number of outputs is shown as a function of the number of inputs for various circuit tolerances in Fig. 11 (j).

On comparing the two systems, the synchronous pair logic is more favourable than the static system. This is primarily due to the fact that the synchronous pair is capable of providing a larger logic gain under practical conditions. This is even more evident in an OR gate where the number of inputs required is only one, and the gate is hence capable of providing a large number of outputs. Two or more phase systems can be used and the considerations outlined in Section 4.4 may be applied.

4.2. *Proposal for an* AND *Gate*

A system of logic that depends on an absence of all inputs for AND gates and on the absence of one input for OR gates might be considered. Such a system need not be markedly dependent on the values of peak current; it does however depend on gating diodes with well defined and consistent characteristics.

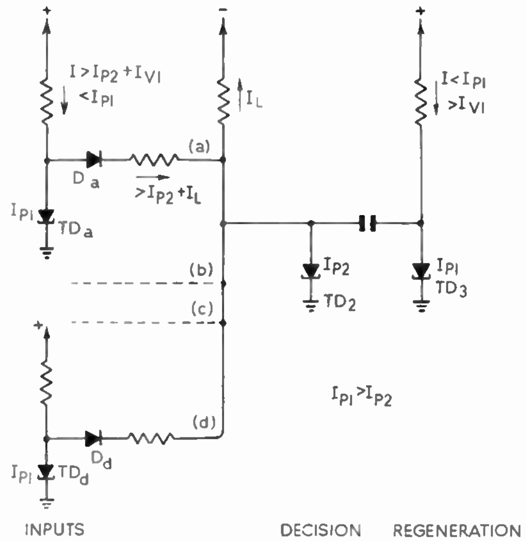


Fig. 11. (k) AND gate when a, b, c, and d are in the 0 state, TD2 switches to the 0 state.

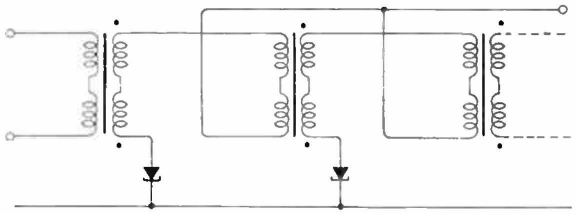
Figure 11 (k) shows a proposal for an AND gate. In the absence of any input, all tunnel diodes are in the "1" or high voltage state. Each input diode, TDa to TDd, is capable of maintaining the "decision", TD2, diode in the "1" state so the peak current value of the decision diode is much less than that of the input diodes. When all the input diodes are in the "0" state, the gating diodes, Da to Dd, are cut-off and a negative bias on the decision diode overcomes the leakage currents of the gating diodes and ensures that the decision diode switches to the "0" state. The decision diode is capacitively coupled to the regeneration diode. The decision diode, on switching, provides enough energy to switch the regeneration diode, TD3, at the same time, even though the regeneration diode has a higher peak current. The regeneration diode can operate more decision diodes and so on. Unfortunately the bias conditions on the regeneration diode are critical. Satisfactory AND and OR gates using an asynchronous system and giving a large fan-in and fan-out have not been devised yet.

4.3. *Divide-by-Two Circuits*

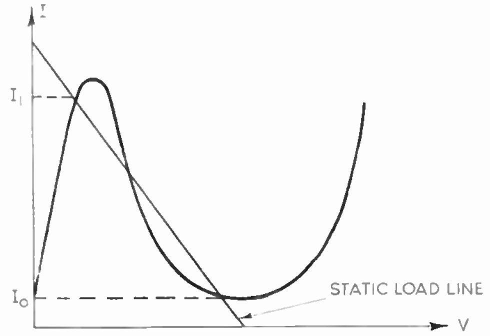
The popularity of the Eccles-Jordan type of circuit in computing systems indicates that there is the need for a similar circuit which would make use of the advantages offered by tunnel diodes.

The low voltages involved make gating for the purpose of routing pulses difficult. The circuit shown in Fig. 12 makes use of the difference between high and low currents obtained, when a diode is biased from a low resistance source. Saturation of a core provides pulse routing.

The turns on opposing transformer windings are adjusted so as to provide a small negative pulse which changes the state of the diode from I_0 to I_1. The higher current saturates the predominant transformer, changing the polarity of the next pulse, which will return the circuit to I_0. The circuit is sensitive to input pulses of only one polarity, making it a simple matter to cascade stages as shown. However, one is limited to low frequencies due to the large number of turns required to saturate the transformers. The transformers can take the form of small ferrite rings.

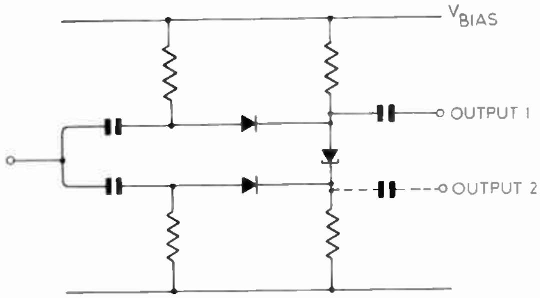


(*a*) Divide-by-two circuit using saturable cores.

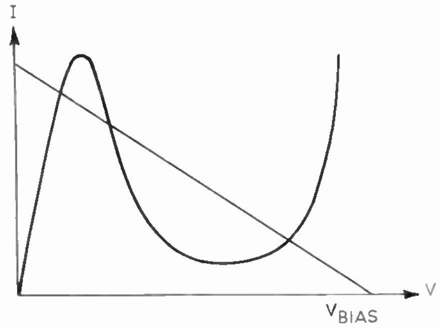


(*b*) Tunnel diode characteristic and loadline.

Fig. 12.



(*a*) Divide-by-two circuit using gating diode.



(*b*) Tunnel diode characteristic and load line.

Fig. 13.

Biasing from a higher resistance source removes some of the stringent requirements placed on the biasing voltage tolerances and is hence a step towards better designability. This and a higher speed are the salient points of the circuit shown in Fig. 13.

Two gold-bonded diodes (S.T.C. DK12) or backward diodes (S.T.C. JK100) are used to produce pulse routing by means of the potential developed across them by the voltage state of the tunnel diode. This circuit provides an output of either phase, each output pulse being equal in amplitude to half the voltage change across the diode, since the system has to be inherently balanced. If a gallium arsenide tunnel diode is used, the gating diodes can be of the gold-bonded type; with germanium tunnel diodes, backward diodes are required.

The division of biasing resistance in the tunnel diode switch makes a pulse regenerating phase splitter possible and this is illustrated in the divide-by-two circuit shown in Fig. 14.

The first tunnel diode (TD1) serves as pulse regenerator and phase splitter providing out-of-phase pulses to the gating diodes. The diodes (which are backward diodes) are again biased for conduction or blocking by the state of the second tunnel diode (TD2) which also provides the output.

It has been mentioned in a previous section that a constant current source is required to trigger the tunnel diode. This is clearly not the case in the circuits of Figs. 13 and 14 and it is necessary to add a

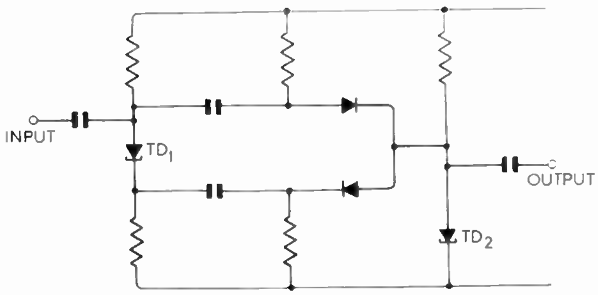


Fig. 14. Divide-by-two circuit using a phase splitter.

small inductance in series with the operative tunnel diode. This is a drawback at higher frequencies. As a rule, it is possible to approach an input repetitive period ten times longer than the minimum theoretical switching time of the diodes. Due to its regenerative nature, the circuit of Fig. 14 is better suited to direct cascading than that of Fig. 13 where some additional inter-stage amplification might be required.

A diode pair, as described in Section 3.1, may be used instead of TD2 in Fig. 14. Such an arrangement is shown in Fig. 15.

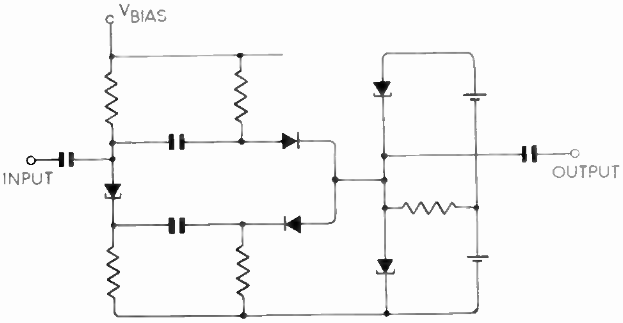


Fig. 15. The circuit of Fig. 14 with a pair of diodes in the output stage.

4.4. *Shift Registers and Ring Counters*

4.4.1. Square clock pulse mode

The descriptions of the systems which follow apply to shift registers but it is understood that the same circuits can be used for ring counters simply by connecting the input to the output and inserting only one pulse. The most reliable way to use tunnel diodes in shift registers and ring counters is in pairs, where the shift pulse is applied directly to the pair and acts also as the bias voltage. Such a circuit is shown in Fig. 16.

The shift pulses are applied in antiphase to the two shift lines. The diode balance is arranged so that if no input is present, the voltage between point P and earth is the low or "0" value. A small trigger current, I_T,

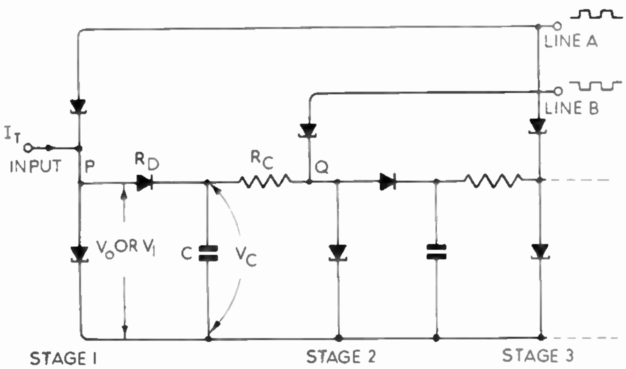


Fig. 16. Two-phase shift register.

ensures that the voltage at P goes to its high, or "1", value. This current can be made as small as 100 μA for $I_P = 5$ mA $\pm$ 2% without putting too stringent a requirement on the matching of the diode peak currents. If a "1" is stored in the first stage, the voltage at P is V_1 and this charges capacitor C to V_C through the coupling diode which is of the gold-bonded variety. As the first stage is de-energized, the charge on C will provide a triggering current to the second stage, at the same time as the clock pulse is applied to it. The circuit follows the following simple d.c. laws:

$$V_C = V_1 \frac{R_C + R_T}{R_{Df} + R_C + R_T}$$

where R_C is the coupling resistance required to build up the charge on capacitor C

R_T is the slope resistance of the tunnel diode near the origin

R_{Df} is the forward resistance of the coupling diode.

Also
$$\frac{V_C}{R_C + R_T'} = I_T > I_{T\min}$$

where $I_{T\min}$ is the minimum trigger current required to change the state of the stage

and R_T' is the tunnel diode resistance at the critical decision point, i.e. when the peak currents of the two diodes are coincident due to the increasing clock voltage.

$$R_T \simeq R_T'$$

hence
$$I_T \simeq V_1/(R_{Df} + R_C + R_T)$$

It should be noted that the voltage developed at P when the stage is in the "0" state (V_0) is low enough to make R_{Df} high, preventing the following stage from being triggered. A similar situation applies when stage 2 is being switched on. At the critical point, 50 mV appears at point Q which is heavily attenuated before it reaches stage 4, hence making it impossible for that stage to switch on.

When Q is at V_1, the ratio between R_C and R_{Db} (the back resistance of the coupling diode) is such as to prevent stage 1 from switching on. Some typical circuit values are given below:

$V_0 = 30$ mV
$V_1 = 350$ mV
$V_C = 130$ mV
$R_C = 300\ \Omega$
$R_{Df} = 500\ \Omega$ (when P is at V_1)
$R_{Db} = 22\,000\ \Omega$
$R_T = 20\ \Omega$
$I_T = 420$ μA
$I_{T\min} = 100$ μA

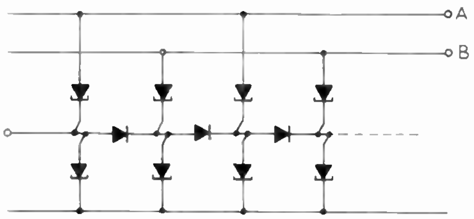
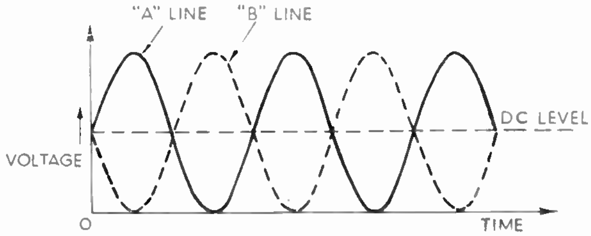


Fig. 17. (*a*) Sinusoidal clock pulse shift register.



(*b*) Clock waveforms.

The value of the capacitance is largely determined by the rise time, fall time and delay between the two shift pulses.

4.4.2. Sinusoidal clock pulse mode

In this mode of operation the shift lines are fed with out-of-phase sinusoidal waveforms superimposed on a d.c. voltage. This system has the advantage of being independent of the clock pulse frequency and employs few components. The circuit with the applied clock pulses is shown in Figs. 17 (*a*) and 17 (*b*).

The relationship between the clock voltage and the tunnel diode pair characteristics is shown in Fig. 18.

The stage is sensitive to an input pulse at t_p, the point of peak current coincidence where the applied voltage is V_{crit}. If the voltage across the lower diodes is considered, it is possible to see how the one stage triggers the next. This is shown in Fig. 19 where the diode junction voltages and two subsequent stages are considered simultaneously.

It is clear that when stage B is at V_{crit}, the voltage at the diode junction of stage A is at V_{trig} and this is enough to trigger stage B.

In practice it is found that a germanium point contact diode is an ideal coupling element, providing reliable operation.

5. The Backward (or Unitunnel) Diode

Figure 20 shows the characteristic of a germanium backward diode. The backward diode is of interest for two reasons:

(1) In the high conduction direction it conducts at a lower voltage than "conventional" diodes with the same junction capacitance. This makes it suitable for a gating device for use with tunnel diodes.

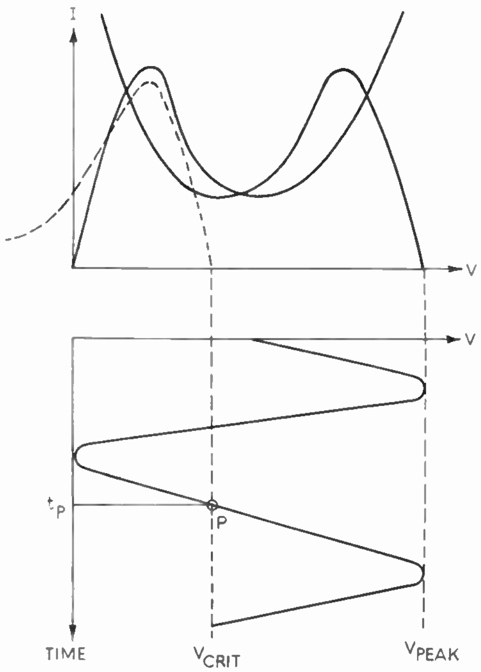


Fig. 18. Clock waveforms with respect to the tunnel diode characteristics.

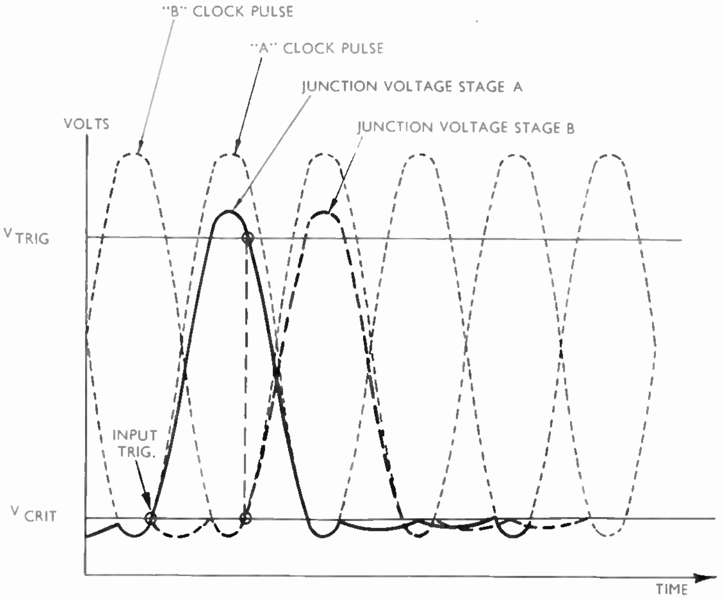


Fig. 19. Clock and tunnel diode waveforms in two consecutive stages.

(2) Its "breakdown" characteristic in the low conductance direction is well-defined and consistent from one device to the next. This makes it suitable as a low level amplitude limiter or discriminator.

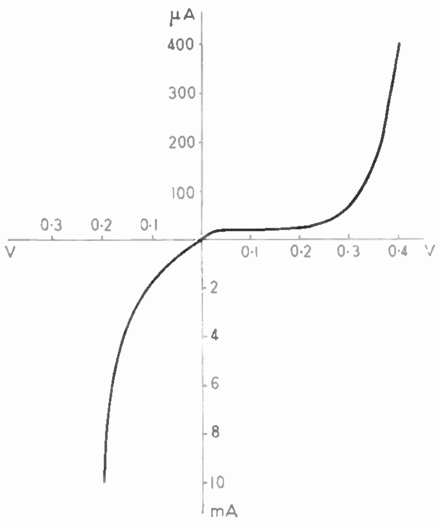


Fig. 20. Backward diode characteristic.

For most practical purposes minority carrier effects in backward diodes can be ignored, and the only factor affecting their high frequency performance is their junction capacitance. The junction capacitance is related to the current that will flow for a given voltage in the low conductance direction. The capacitance of a diode having the characteristics shown in Fig. 20 might be about 20 pF when it is biased at 0·2 volts in the low conduction direction.

Backward diodes should be considered for all gating functions in tunnel diode switching circuits, especially in circuits containing germanium tunnel diodes. They may also be considered in transistor switching circuits where they are capable of performing gating functions at the voltage levels associated with the base-emitter circuit rather than the collector circuit. With transistors, gallium arsenide backward diodes may be more suitable than germanium diodes.

6. Gallium Arsenide and Germanium as Materials for Tunnel Diodes in Switching Circuits

Gallium arsenide has the advantage of giving a greater voltage output making problems of differential conduction in gating diodes much less severe. It also shows promise of giving a much higher peak-to-valley ratio and this may be of importance in many circuits. A comparison between the speed of devices made from the two materials is not possible at this stage of development as the ultimate speed may depend more on fabrication problems than on the inherent properties of the materials concerned. (Other things being equal, the greater voltage swing of the gallium arsenide is a disadvantage as far as speed is concerned.)

When used with transistors, it is probable that the greater voltage swing of gallium arsenide will be an advantage although there are applications where germanium tunnel diode can be used satisfactorily with germanium transistors.

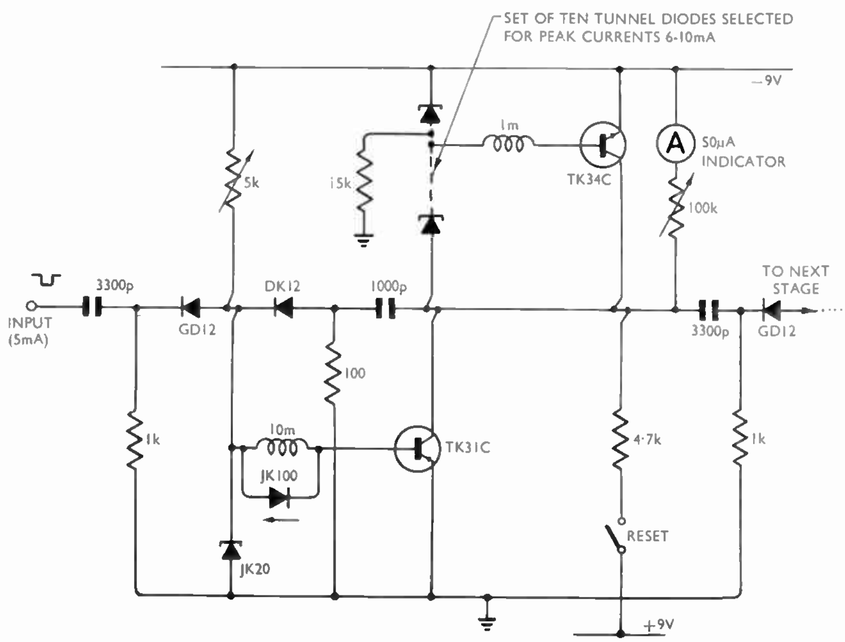


Fig. 21. One stage of a pulse counter.

The fact that gallium arsenide will work at much higher temperatures is to some extent off-set by the fact that it will naturally dissipate more power than a germanium device in the high voltage state. Some gallium arsenide diodes have proved to be unreliable in service and it is unlikely that they will be widely adopted until the reliability problem is solved.

7. A Pulse Counter and Digital Multiplier

The circuit shown in Fig. 9 can be used to make a simple decade pulse counter. The method of cascading stages is shown in Fig. 21. The tunnel diodes used were specially selected for the purpose. They have peak currents ranging from 6 to 10 mA. Using the components indicated, the circuit can be made to count pulses which have a repetitive frequency of up to 1 Mc/s. The input pulse should be a negative 5 mA current with a maximum width of 10 μs. The input impedance of the circuit is chiefly dictated by the input diode and can be as low as 30 Ω.

The scaling system can be used in a simple digital multiplying machine the block diagram of which is shown in Fig. 22. The machine is capable of multiplying two three-digit numbers and exhibiting the product on decade meters. The shifting and operating actions are performed by a manually operated rotary switch.

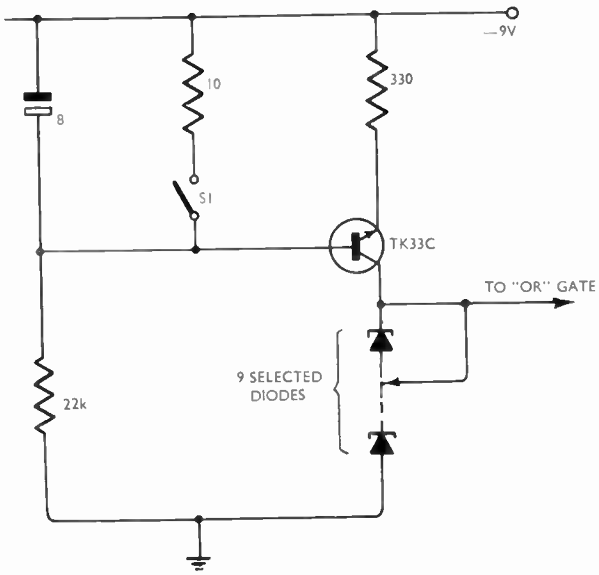


Fig. 23. The "operator" circuit.

The "operator" circuit (Fig. 23) provides a ramp of current to the tunnel diode chain in which the diodes are selected to have differing peak currents. As each diode changes state, it provides an output pulse. Thus the number of output pulses is determined by the number of diodes in the chain, selected by means of a decade switch. The pulses thus produced are fed through one stage of amplification to an OR gate as shown in Fig. 24. A pulse appearing at the OR gate triggers the 45 mA tunnel diode which supplies a ramp of current to the tunnel chains in the three "operand" circuits. Each of these circuits again produces a number of pulses determined by the selection of a number of diodes in the chain. Thus multiplication is achieved. The output pulses from the "operand" circuit have to be amplified by an emitter-follower circuit (not shown) before being fed to the scalers through the rotary "shift" switch.

Delay elements have to be used between the scalers so as to avoid coincidence of input and overflow pulses from the previous stage. These can take the form of a simple Eccles-Jordan monostable multivibrator.

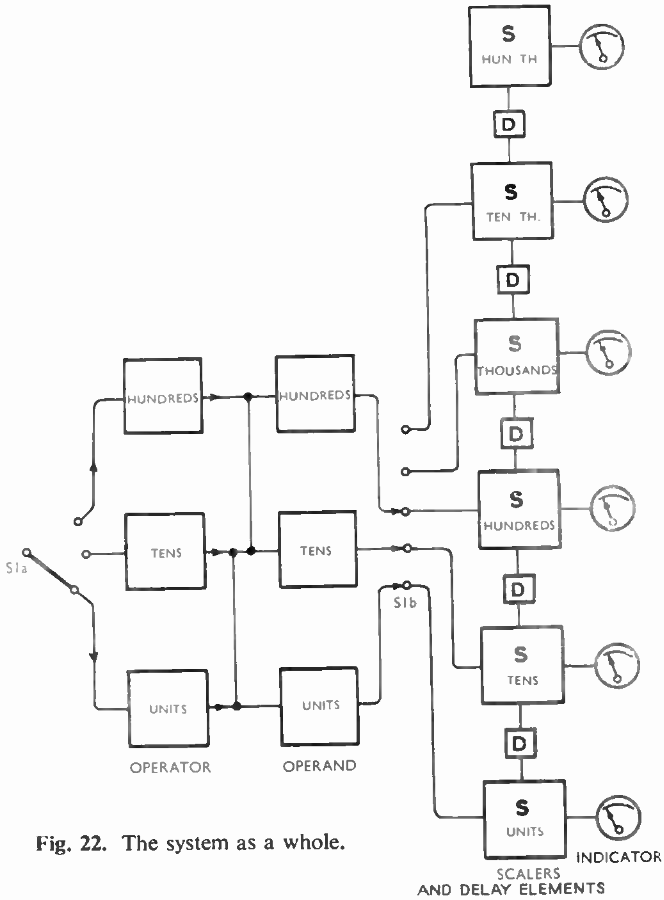


Fig. 22. The system as a whole.

8. Conclusions

The tunnel diode is capable of performing many of the functions normally performed by other active semiconductor elements, but there are difficulties associated with driving one circuit from another. These difficulties can generally be overcome but only

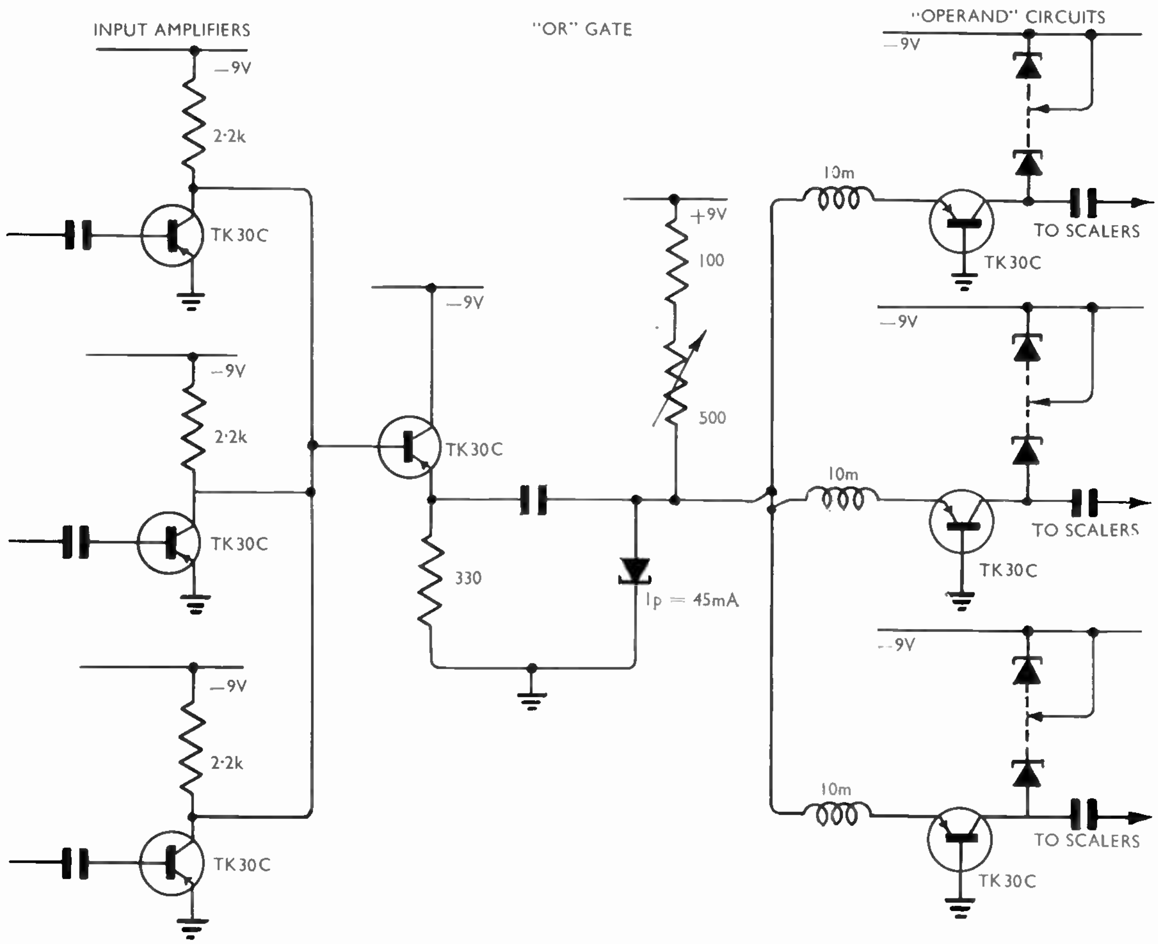


Fig. 24. The OR gate and "operand" circuits.

at the expense of increased complexity. We have seen that tunnel diodes are to a large extent complementary to transistors and their use in transistor circuits may be advantageous, both in terms of performance and cost. The much higher sensitivity of a tunnel diode pair in synchronous logic, as opposed to asynchronous logic, makes possible a computing system based almost entirely on tunnel diodes.

It is clear that much work remains to be done, both in devising circuits using tunnel devices, and in developing devices that depend for their operation on the tunnel effect. We are at the beginning of a new and exciting chapter in the history of semiconductor development.

9. Acknowledgments

The authors would like to thank Standard Telephones and Cables Limited for permission to publish this paper and to acknowledge the contribution of their colleagues to the work on which it is based.

10. Appendix 1

Static Logic Calculations

Let α be the tolerance on the trigger current I,

i.e. trigger current $= I(1 \pm \alpha)$

Let β be the tolerance on the peak current of the diode,

i.e. peak current $= I_P(1 \pm \beta)$

Let n be the number of inputs required to trigger the diode from a low voltage state (0) to a high voltage state (1).

Hence $nI(1-\alpha) > I_P(1+\beta)$

Therefore $$n > \frac{I_P}{I}\left(\frac{1+\beta}{1-\alpha}\right) \qquad \text{......(4)}$$

$n-1$ pulses must not trigger the gate hence:

$$(n-1)I(1+\alpha) < I_P(1-\beta)$$

Thus for a real value of n

$$\frac{I_P}{I}\left(\frac{1-\beta}{1+\alpha}\right) + 1 > \frac{I_P}{I}\left(\frac{1+\beta}{1-\alpha}\right)$$

Therefore $\frac{I_P}{I}\left[\left(\frac{1-\beta}{1+\alpha}\right)-\left(\frac{1+\beta}{1-\alpha}\right)\right] > -1$

i.e. $I > 2I_P\left(\frac{\alpha+\beta}{1-\alpha^2}\right)$

In the limit condition, we have a minimum value of trigger current I_{min}, where

$$I_{min} = 2I_P\left(\frac{\alpha+\beta}{1-\alpha^2}\right) \qquad \ldots\ldots(5)$$

This yields a maximum value of n given by substituting in (4) above:

$$n_{max} = \frac{(1+\alpha)(1+\beta)}{2(\alpha+\beta)} \qquad \ldots\ldots(6)$$

If m is the number of outputs and the output current equals the trigger current

$$m = \frac{I_P(1-\beta) - \frac{I_P}{A_{min}}(1+\beta)}{I(1+\alpha)}$$

where A_{min} is the minimum peak-to-valley ratio. And from eqn. (5)

$$m = \frac{(1-\alpha)\left[(1-\beta) - \frac{(1+\beta)}{A_{min}}\right]}{2(\alpha+\beta)}$$

$$= n_{max} - 1 - \frac{(1+\beta)}{2A_{min}(\alpha+\beta)} \qquad \ldots\ldots(7)$$

11. Appendix 2

Threshold Calculations for Tunnel Diode Pair Logic

Let

α be the tolerance on the trigger current I

β be the tolerance on the peak tunnel diode currents I_{PH} and I_{PL},

where I_{PH} is the higher peak current

I_{PL} is the lower peak current.

Considering an AND gate where the number of inputs required to trigger the gate is n, i.e. $n-1$ inputs must not trigger the gate:

$$nI(1-\alpha) > I_{PH}(1+\beta) - I_{PL}(1-\beta)$$

$$(n-1)I(1+\alpha) < I_{PH}(1-\beta) - I_{PL}(1+\beta)$$

$$n > \frac{I_{PH}(1+\beta)}{I(1-\alpha)} - \frac{I_{PL}(1-\beta)}{I(1-\alpha)}$$

also $$n < \frac{I_{PH}(1-\beta)}{I(1+\alpha)} - \frac{I_{PL}(1+\beta)}{I(1+\alpha)} + 1$$

The limit for a real value of n is given by

$$\frac{I_{PH}(1+\beta)}{I(1-\alpha)} - \frac{I_{PL}(1-\beta)}{I(1-\alpha)} = \frac{I_{PH}(1-\beta)}{I(1+\alpha)} - \frac{I_{PL}(1+\beta)}{I(1+\alpha)} + 1$$

Therefore

$$I_{min} = I_{PH}\left[\frac{1+\beta}{1-\alpha} - \frac{1-\beta}{1+\alpha}\right] + I_{PL}\left[\frac{1+\beta}{1+\alpha} - \frac{1-\beta}{1-\alpha}\right]$$

$$= \frac{1}{1-\alpha^2}[I_{PH}(1+\alpha+\beta+\alpha\beta-1+\beta+\alpha-\alpha\beta) + I_{PL}(1+\beta-\alpha-\alpha\beta-1+\beta-\alpha+\alpha\beta)]$$

$$= \frac{2}{1-\alpha^2}[I_{PH}(\alpha+\beta) - I_{PL}(\alpha-\beta)] \qquad \ldots\ldots(8)$$

The maximum value of $n(n_{max})$ is given by

$$n_{max} - 1 = \frac{I_{PH}}{I_{min}}\left(\frac{1-\beta}{1+\alpha}\right) - \frac{I_{PL}}{I_{min}}\left(\frac{1+\beta}{1+\alpha}\right)$$

$$= \frac{1-\alpha}{2}\left[\frac{I_{PH}(1-\beta) - I_{PL}(1+\beta)}{I_{PH}(\alpha+\beta) - I_{PL}(\alpha-\beta)}\right]$$

Therefore

$$n_{max} = \frac{(1+\alpha)}{2}\left[\frac{I_{PH}(1+\beta) - I_{PL}(1-\beta)}{I_{PH}(\alpha+\beta) - I_{PL}(\alpha-\beta)}\right] \qquad \ldots\ldots(9)$$

Let now the number of outputs be m

Hence $$m = \frac{I_{PL}(1-\beta) - I_V^*}{I(1+\alpha)}$$

where I_V^* is the highest possible valley current of the high peak current tunnel diode.

Therefore $$I_V^* = \frac{I_{PH}(1+\beta)}{A_{min}}$$

where A_{min} is the minimum specified current ratio

Hence $$m = \frac{I_{PL}(1-\beta) - \frac{I_{PH}}{A_{min}}(1+\beta)}{I(1+\alpha)}$$

Substituting I_{min} of eqn. (8) for I

$$m = \frac{(1-\alpha)}{2}\left[\frac{I_{PL}(1-\beta) - \frac{I_{PH}}{A_{min}}(1+\beta)}{I_{PH}(\alpha+\beta) - I_{PL}(\alpha-\beta)}\right]$$

hence

$$\frac{m}{n} = \left(\frac{1-\alpha}{1+\alpha}\right)\left[\frac{I_{PL}(1-\beta) - \frac{I_{PH}}{A_{min}}(1+\beta)}{I_{PH}(1+\beta) - I_{PL}(1-\beta)}\right] \qquad \ldots\ldots(10)$$

Manuscript first received by the Institution on 19th May, 1961 and in final form on 30th December, 1961. *(Paper No. 709.)*

PROCEEDINGS OF THE COUNCIL

Much of the business of the Council is concerned with reports from Standing Committees, whose activities are featured in the Journal. *A digest of other matters of immediate and particular interest to all members will be given in these notes from time to time.*

The third meeting of the Council of the Institution since the grant of the Royal Charter of Incorporation was held on 8th February, 1962. The Chair was taken by the President, Admiral of the Fleet The Rt. Hon. Earl Mountbatten of Burma, K.G.; 25 members of Council were present.

Institution Meetings.—Reference was made by the President to National Productivity Year. Lord Mountbatten proposed that the Institution's 1963 Convention should have "Electronics in Productivity" or a similar title as its main theme, and it was agreed that a Convention Committee be appointed to make arrangements.

The North Western Section Committee placed before Council proposals for Institution activities during the British Association's meeting in Manchester in September 1962. It was agreed that the Institution should sponsor papers and that the associated Science Fair for young people would also be supported.

Proposals for Research Activities.—The President recommended that in view of the importance of encouraging radio and electronics research it was desirable that the Council should have a permanent standing "Research Committee". The Council appointed the first members to serve upon this Committee (listed on page 162). Its immediate task will be to consider the establishment by the Institution of research grants and scholarships for work on radio and electronics at universities and to review present research facilities in Universities, Government Departments and Industry.

Education Matters.—The Education and Training Committee reported invitations to appoint Institution representatives to serve on the following bodies:

The Engineering and Science Advisory Committee of Southampton Technical Colleges (*Mr. W. Renwick, M.A., B.Sc., (Member) was appointed*).

The Electrical Engineering Advisory Committee of the College of Technology, Barnsley (*Mr. D. Shaw, B.Sc., (Associate Member) was appointed*).

The Advisory Committee for Electrical Engineering (including Radio and Telecommunications) of the Regional Advisory Council for Technological Education—London and Home Counties (*Professor E. E. Zepler, Ph.D., (Member) was appointed*).

Joint I.E.E.–Brit.I.R.E. Committee.—A report was received from the Institution's delegates to the Joint Committee, Messrs. L. H. Bedford, W. E. Miller and Colonel G. W. Raby, and Mr. I. Maddock, on the first meeting of the Committee held on 6th February.

Council approved the suggestion that the Executive Committee should prepare a memorandum outlining the Institution's policy in developing its activities, and that the memorandum should be placed before the Joint Committee immediately.

The Institution Overseas.—The Council approved the initiation of steps to secure incorporation of the Institution in Canada.

After the conclusion of the Council meeting the President entertained members to dinner. During the course of the evening Admiral of the Fleet the Earl Mountbatten of Burma, on behalf of all the Officers and Council, presented the Secretary, Mr. Graham D. Clifford, with an illuminated address to express thanks for Mr. Clifford's personal help in securing the Institution's Charter. The President also commented on the fact that Mr. Clifford had completed 25 years of service to the Institution.

International Television Conference, 31st May–7th June, 1962

The Conference on Television which is being organized by the Electronics and Communications Section of The Institution of Electrical Engineers in association with the Television Society,† marks an important milestone in the history of television, since it follows the 25th anniversary‡ of the start by the British Broadcasting Corporation of the first public television service in the world.

As noted on page 162, the arrangements already made for this Conference were discussed at the first meeting of the newly-appointed "Institution of Electrical Engineers and British Institution of Radio Engineers Joint Committee". The Committee considered an invitation from the I.E.E. to the Brit.I.R.E. to participate in the Conference, but since arrangements were now well advanced it was agreed that it was too late for the Brit.I.R.E. to take a complete part in the planning of the programme. Nevertheless, the participation of Brit.I.R.E. members in the Conference was welcomed and subsequently the Council of the I.E.E. have been pleased to direct that members of the Brit.I.R.E. should pay the same registration fee as members of the I.E.E.

The Conference will deal with technological aspects of the whole television field, including programme origination, transmission, reception, standards, colour, the use of television both for industrial and educational purposes, pay television, mobile and fixed links, including the use of communication satellites.

The period chosen for the Conference will overlap the last three days of the International Instruments, Electronics and Automation Exhibition, which is to be held at Olympia, London, at that time and at which most British and many oversea manufacturers of professional television equipment and apparatus will be exhibiting. Arrangements are being made for those attending the Conference to visit the Exhibition and special transport will be available from the Conference Halls to the Exhibition building.

The provisional programme of the Conference is given below. Associated with the technical programme will be opportunities for those taking part in the Conference to visit research and industrial establishments concerned with television.

Further particulars about the Conference are available on application to the Secretary, The Institution of Electrical Engineers, Savoy Place, London, W.C.2. Members of the Brit.I.R.E. should quote their membership when applying for registration.

PROGRAMME OF SESSIONS

Thursday, 31st May

Opening Session

Friday, 1st June

MORNING: "Systems Standards"

AFTERNOON: "Signal Pick-up Tubes"

Monday, 4th June

MORNING: "Frequency Assignment Problems"
"Television Recording—Programme and Commercial"

AFTERNOON: "Standards Conversion"
"Medical Uses"

Tuesday, 5th June

MORNING: "Point-to-Point Links"
"OB Equipment"

AFTERNOON: "Studio Design"
"Scientific Applications"
"Data Transmission"

Wednesday, 6th June

MORNING: "Wire Broadcasting for Domestic Television" "Studio Design and Lighting" "Industrial Applications"

AFTERNOON: "Broadcast Transmitter Equipment" "Low Light Level Applications"

As indicated in the above programme, there will be simultaneous sessions on three days of the conference. Full information as to detailed arrangements is available direct from the Institution of Electrical Engineers.

† *J. Instn Elect. Engrs*, **7**, page 470, July 1961. ‡ "Twenty-five years of television", *J. Brit.I.R.E.*, **22**, page 526, December 1961.

U.D.C. 621.311.6

On the Design of Small Economical Radio Frequency E.H.T. Supplies

By

J. K. MOORE (*Associate*)†

Summary: The design and construction of a stabilized r.f. e.h.t. unit is described. The advantages over previously published designs are in smaller size and more economical construction. The design data are applicable to all cases where a low current high voltage e.h.t. supply is required at a minimum cost and size.

List of Symbols

K_c critical coupling
Q_1 magnification factor of coil number 1, etc.
Z_1 impedance of coil number 1, etc.
n number of rectifiers in cascade
L inductance of coil in henries
$\bar{R}$ d.c. voltage/d.c. current
ω $2\pi f$
$\hat{E}_2$ peak value of voltage
E_c electrical stress in a cavity
ε permittivity of insulating material
ρ resistivity of insulant
V_i discharge inception voltage
EG electrical strength of a gas
r d.c. resistance of coil
C self-capacitance of coil
f frequency

1. Introduction

The general requirements of an e.h.t. supply (not necessarily listed in the order of importance) are:

1. Minimum volume.
2. Low power input.
3. Economical in manufacture.
4. Reliability in use.
5. Capability of delivering several differing high voltages simultaneously.
6. Non-lethal.

Items 1 and 3 indicate the advantage of a transformer working at high frequencies, as this would be less bulky than a 50 c/s transformer stepping up the mains voltage. Also the use of high frequencies allows a considerable reduction in the size and cost of the high voltage smoothing capacitors. Difficulties which arise through using a high frequency source of supply are: load on the general h.t. supplies, stabilization of the high voltage output due to the high source impedance, and the design of a suitably small high voltage, high frequency transformer.

The design of a high-frequency transformer is by far the greatest difficulty. A number of excellent designs[1, 2] have been published utilizing air-cored transformers and a design for a ferrite cored transformer is also available.[3]

In order to understand the problem, a brief review of the design and electrical properties of air-cored transformers will be helpful.

A high-voltage radio-frequency transformer is a particular case of a pair of tuned coupled circuits.

If critical coupling is used then

$$K_c = 1/Q_1 Q_2 \qquad \ldots\ldots(1)$$

giving maximum voltage step-up and

$$\hat{E}_2/\hat{E}_1 = \sqrt{\frac{Z_2}{Z_1}} \qquad \ldots\ldots(2)$$

However, critical coupling depends upon secondary Q (Q_2 in eqn. (1)) which will vary with secondary load, hence critical coupling could not be maintained with a varying load, and this would make the step-up ratio vary, giving poor secondary voltage regulation. Further, when critical coupling is used, the secondary power is one-half the total power input to the transformer and efficiency cannot exceed 50%; if less than critical coupling were used efficiency would be even less.

It is clear that for improved secondary voltage stability and good transformer efficiency the coupling co-efficient should be much greater than critical value.

Additional practical difficulties are that as wave-winding is generally used in order to withstand voltage stresses between layers it is difficult to obtain a close coupling due to the physical distance between the primary and secondary windings.

† Formerly Marconi Instruments Ltd., St. Albans, Hertfordshire; now with De Havilland Aircraft Co. Ltd., Hatfield, Hertfordshire.

2. Ferrite-Cored Transformers

By the use of a suitable ferrite core many of the above difficulties can be overcome and the physical size of the transformer can be considerably reduced.

It is desirable to work at a high frequency so that smoothing of the rectified high voltage output may be obtained by low values of capacitors which are both small and economical. Also the total energy stored will be relatively low and therefore non-lethal and should accidental overload occur the excessive damping on the driving source will reduce oscillations and thus severely curtail the output power. By working at frequencies over 40 kc/s one of the disadvantages of Ferroxcube, magnetostrictive noise, is avoided.

3. Effective Rectifier Loading

As the oscillator frequency will generally be greater than 40 kc/s valve rectifiers will have to be used as metal rectifiers do not function satisfactorily at these frequencies, therefore additional power will be required for the valve heaters. In any case valves are preferable to metal rectifiers for voltages above approximately 1·5 kV as they are cheaper, smaller, and have a lower self-capacitance.

For satisfactory operation the secondary resonant circuit should have an impedance which is not less than ten times the impedance of the load presented to it.[1, 2]

Thus
$$Z_0 = \not< 10 R_L \qquad \ldots\ldots(3)$$

R_L obeys the law

$$R_L = \frac{\bar{R}}{2n^2} \qquad \ldots\ldots(4)$$

where
$$\bar{R} = \frac{\text{d.c. output voltage}}{\text{d.c. output current}} \qquad \ldots\ldots(5)$$

and n is the number of rectifier stages in cascade.

Therefore we can see that the impedance R_L presented to the secondary is reduced by $\frac{1}{2}$, $\frac{1}{8}$ and $\frac{1}{18}$ for half wave, doubler, and tripler respectively.

It will be noted that although the use of cascaded rectifiers reduces the required impedance of the secondary winding it does not alter the total power this winding has to supply.

Further advantages of using voltage multiplication are the reduction of total turns required and consequently the self-capacitance of the coils. These must of course be weighed against the increase of costs, etc., due to the increased number of rectifiers and capacitors.

Factors governing secondary resonant impedance are

$$Q = \frac{\omega L}{r} \qquad \ldots\ldots(6)$$

$$= \frac{1}{\omega C r} \qquad \ldots\ldots(7)$$

$$Z = Q\omega L \qquad \ldots\ldots(8)$$

The secondary power loss

$$P_2 = \frac{E_2^2}{Z_2} \qquad \ldots\ldots(9)$$

The equivalent shunt loss resistance is given by

$$Z_2 = \frac{L}{rC} \qquad \ldots\ldots(10)$$

Hence from eqns. (9) and (10) we have

$$P_2 = \frac{E_2^2 rC}{L} \qquad \ldots\ldots(11)$$

where r = coil d.c. resistance and C = self capacitance of coil.

Thus for high secondary resonant impedance,

$$Z_2 = \frac{L}{rC}$$

the transformer must have

(1) low secondary resistance,

(2) low secondary capacitance,

(3) high secondary inductance.

For (1) we can make the length of the mean turn as short as possible, and have as few turns as possible. We can assume the a.c. resistance of the secondary to be similar to the measured d.c. resistance[3] up to approximately 150 kc/s.

For (2), when the transformer is layer wound it is desirable to have many layers of narrow width with a good spacing between layers[4]

Table 1

Valve	Heater Voltage	Heater Current (amps)	Total Heater Power (watts)	Factor increase on U37	Base
U37	1·4	0·140	0·916	—	Wire ended
U43, EY51, EY86, R12	6·3	0·090	0·567	2·9	Wire ended
U45	6·3	0·120	0·755	3·8	Wire ended
U47	2·0	0·200	0·400	2·0	
DY86	1·4	0·55	0·77	3·8	
R10	4·0	0·500	2·0	10	
IT2/R16	1·4	0·140	0·196	—	
R19	1·25	0·200	0·250	1·25	B9A
R20	2·0	0·350	0·700	3·6	B9A

Considering (3); the secondary inductance is governed by the total turns, the Q factor of the coil and the effective gap in the core.

In units of the type shown it is desirable to run the high voltage heaters from suitably insulated turns around the transformer. Thus the transformer has also to supply the heater power for the valves in addition to the other load requirements, so the advantages of low heater power valves may be quite an important factor. In the design shown U37 valves have been utilized and consume a total power of 1 watt.

Table 1 lists valves which may be used, and shows the relative increased heater power requirements for other valve types in comparison with the U37.

It was decided that the choice of turns per volt should be such that an integral number of turns provided the correct heater voltage for the U37 valves used, and this number of turns should be kept to a minimum in order to keep the physical size of the transformer as small as possible. If this arrangement is not possible resistors may be used to ensure the heater voltages are correct.

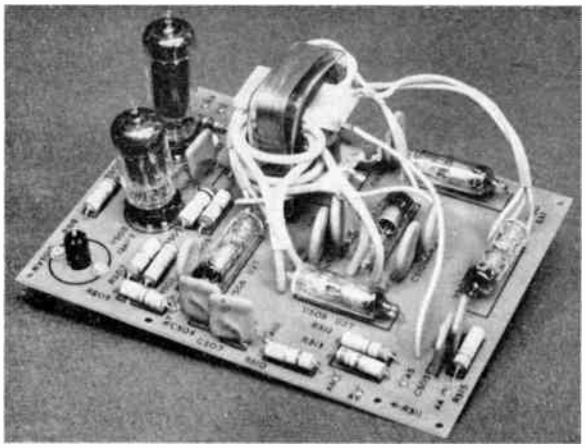


Fig. 1. This unit supplies −1250 V d.c. stabilized, a floating supply of −1520 V d.c., and a positive supply of 8·75 kV d.c.

As may be seen from Fig. 1, two turns around the outer limbs of the "E" core are used for heaters. This allows a higher voltage per turn to be generated in the centre core, which in turn requires fewer turns to obtain the required high voltage output and so assists in obtaining a high frequency of operation.

4. Insulation

The electric strength of insulation is extremely difficult to determine. It depends upon such factors as thickness, homogeneity, the presence of gaseous inclusions, moisture, the physical shape of conductors adjacent to the insulation, ambient temperature, humidity, pressure, and the frequency and waveform of the applied voltages. Clearly unless the test conditions are stated the figure of electric strength of a material is meaningless.[5] However, in small high-frequency transformers for electronic equipment certain of these phenomena have considerably greater importance.

A most important consideration is to avoid internal discharges within the transformer. Discharges occur mainly when gas-filled cavities are present in the transformer insulation, and are due to concentrations of stress across the cavity, a region of lower permittivity than the surrounding insulation. The electrical stress E_c in a cavity is greater than the stress in the surrounding insulation, E, by an amount ε, the permittivity of the solid insulation. If the permittivity of the gas is taken as unity and the cavity is taken to be a thin wide disc normal to the electric field,

$$E_c = \varepsilon E \qquad \ldots\ldots(12)$$

The discharge inception level, that is, the voltage required to initiate discharges across the cavity, is given by[10]

$$V_i = \frac{EG(t + t'(\varepsilon - 1))}{\varepsilon} \qquad \ldots\ldots(13)$$

where EG is the electric strength of the gas at a pressure p, the cavity is of thickness t' surrounded by insulation of thickness t.

If one discharge only occurred near the peaks of each cycle we would see that discharges are proportional to frequency.

If, however, the stressing voltage was d.c. a discharge would occur as the applied voltage reached the discharge inception level, V_i. If this voltage is maintained at this level further discharges will only occur at intervals determined by the rate of recovery of the voltage across the cavity after a discharge. If there is negligible surface leakage within the cavity, which is generally the case with good quality insulants, the recovery time of the cavity will approximate to

$$t_r = \varepsilon \rho \times 10^{-13} \text{ seconds} \qquad \ldots\ldots(14)$$

where ρ is the resistivity of the insulant in ohm-cm and ε is its relative permittivity. If $\rho = 10^{12}$ ohm-cm, as in poor quality insulation, the recovery time is rather less than one second and fairly frequent discharges will occur. However, in high quality insulation t_r may be several hours or days. It is thus clear that stresses in insulation subjected to alternating voltages are considerably more critical in their design than when direct voltages are considered. It can also be seen that the breakdown potential of an insulant, when expressed in volts/mil, has very little meaning when the insulant is used under conditions of a.c. stresses, as compared to use under d.c. stresses.

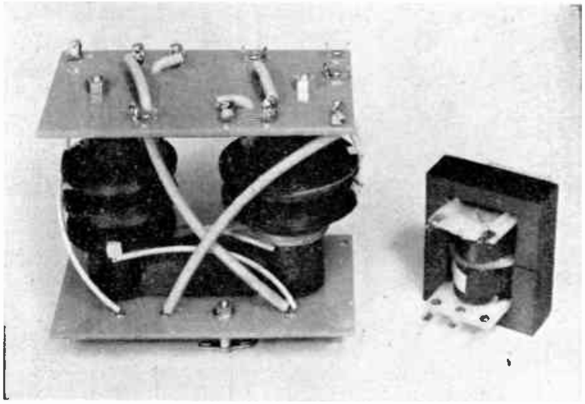


Fig. 2. (*a*) The possible reduction in size obtainable by layer winding as compared to conventional wave winding.

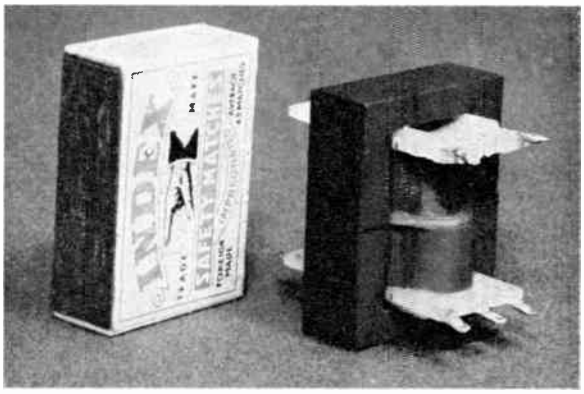


Fig. 2. (*b*) Relative size of transformer described.

For optimum design the insulation should have the following characteristics:

(*a*) As free as possible from gaseous inclusions.

(*b*) Of low permittivity to give a high discharge inception stress.

(*c*) Homogeneous, to avoid stress concentrations at localized points.

(*d*) Have low dielectric loss angle, low electrical conductivity, and high thermal conductivity to remove local heat concentrations.

(*e*) Low moisture absorption.

(*f*) Free from surface tracking.

It is also of advantage to use wire that is insulated with polyvinyl acetal (p.v.a.) enamel. This has greater abrasion resistance and flexibility etc., than the cheaper oleo-resinous enamels.

It is useful to know that as the life of insulation is inversely proportional to frequency under certain conditions, such as when discharges are occurring, accelerated life tests may be made on insulants used at low frequency by stressing them with high-frequency discharge-free voltages.

Discharges are rather difficult to measure without specialized apparatus and the most satisfactory method is to use commercial equipment. A suitable equipment has been designed[6] and is available. Briefly a band-pass amplifier with a centre frequency of 100 kc/s and approximately 10 kc/s bandwidth is connected to the specimen under test and a variable high voltage at 50 c/s stresses the specimen. By filtering this supply and the rejecting of the 50 c/s voltage by the band-pass amplifier, any output from the amplifier will be due to the generation of discharges in the specimen under test. As the ionizing time of the discharges in insulation is generally less than 0·1 μs, the amplifier will respond to the transients generated. These are displayed upon an oscilloscope and compared with a pulse of known variable amplitude and a measure of the discharge is thus made. As the capacitance of the specimen may be measured the magnitude of the discharge power may be calculated.

For a more exhaustive survey of contemporary practice a study of reference 5 is suggested. This gives a clear exposition of the problems involved and includes more than 130 references.

5. Practical Transformer—Method of Winding

In order to reduce the size of the transformer, layer winding is used instead of the more conventional wave winding. The improvement obtainable may be seen in Fig. 2(*a*) while Fig. 2(*b*) gives an indication of size.

With wavewinding, the limiting factor in reducing the height of the pile when working at a high volts/turn becomes the breakdown due to discharges between the layers. Due to the nature of winding of such a coil, the worst case, where two vertically adjacent wires have twice the maximum volts per layer across them, occurs continuously throughout the winding. In layer construction the worst condition occurs only at the end of each layer, as illustrated in Fig. 3.

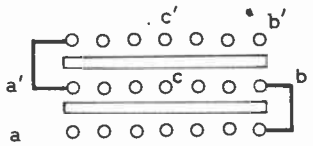


Fig. 3. Voltage differences with layer winding.

The maximum stress per layer is between a and a′ and b and b′. The stress between c and c′ is only one half the worst case. Thus reliability of the transformer as compared to a similarly stressed wavewinding is considerably improved.

The figures in Table 2 of discharge inception voltages were taken on transformers which employed conventional layer winding and had been vacuum impregnated in liquid polythene.

Table 2

Insulation Thickness (mils)	Discharge Inception Voltage (kVr.m.s.)	Discharge Extinction Voltage (kVr.m.s.)
16	1·6	1·2
10	1·2	1·0
6	1·0	0·8

It will be seen from these figures that the discharge inception voltage is greatest for the thickest insulation and that the discharge extinction voltage is less than the inception voltage. This illustrates the desirability of working any insulation at as low a stress as practical, for in a maximum stress design discharges may be initiated due to a sudden surge and then continue at the working voltage.

The former for the winding is injection moulded from nylon type B100M which has a low moisture absorption, low permittivity and loss angle, and high volume resistivity. It is robust in use and economical in manufacture.

The dimensions of the former are suitable for use with Mullard Ferroxcube pieces type FX 1817. These are high permeability A4 grade, for which data are readily available.[7]

6. Practical Transformer—Calculation of Turns

Calculation of turns starts with consideration of the rectifier heater supply. In the first trial transformer, an endeavour should be made to use one turn around the outer limb of an "E" core to provide correct filament voltage of the U37 valves utilized. This valve has 1·4 V heaters taking 140 mA of current. If there is $1{\cdot}4 \times \sqrt{2}$ peak volts generated in the outer limb there should be twice this in the centre core. Hence it is required to generate $2 \times 1{\cdot}4 \times \sqrt{2}$ volts peak per turn.

To obtain -1250 V output for the stabilized winding it would require 326 turns. Similarly using a tripling circuit to obtain the positive supply it is required to generate 8750/3 = 2920 V, thus 2950/4 = 730 turns are required.

In the circuit shown it will be practical to drive the anode of the power tetrode close to the knee of the anode characteristic and with the value shown this will be a swing of approximately 250 V. This represents 60 turns for the anode winding. Empirical tests were made to find an optimum anode/grid turns ratio and this was found to be 3 : 1, giving 20 turns for the grid winding. This agrees with other oscillator data.[9]

Also on the secondary 2 winding it is required to generate -1520 V for a floating winding which is accurately compared to the -1250 stabilized winding to provide control of brilliance for a cathode-ray tube.

In the design shown it was desired to operate the transformer at as high a frequency as possible so a gap was inserted into the transformer core, increasing the frequency and efficiency but the available heater voltage was reduced. Hence the heater turns were increased to two turns per valve and the turns for the other windings were slightly adjusted to suitable values. A suitable transformer was obtained after a few empirical designs had been tried, the principal difficulty in this instance being the high accuracy of balance required between the -1250 V and the -1520 V.

Where the accuracy requirements are not so onerous the initial design on a turns ratio basis should prove satisfactory and the heater voltage may be adjusted, if required, by the use of suitable small resistors in series or shunt.

The whole winding is carefully vacuum impregnated in liquid polyethylene. Transformers with stresses of up to 64 V per mil have been tested for several thousand hours and have been found satisfactory.

Table 3

Transformer Data

Primary anode winding: 60 turns.

Primary grid winding: 20 turns.

Secondary 1 winding: 800 turns tapped at 380 turns.

Secondary 2 winding: 577 turns.

Heater winding: 2 turns each.

The operating frequency is approximately 100 kc/s.

Insulation between layers 0·016 in. of paper, stressed at 40 V per mil.

An insulated spacer is inserted between the cores to give a gap of 0·030 in.

Output voltages (of complete unit).

-1250 V stabilized

-1520 V

$+8750$ V

One of these units was allowed during six months of continuous running to accumulate a very heavy covering of dust particles, and it is interesting to note that the heavy covering of particles did not have any detrimental effects on the printed circuit high voltage board. The printed board is stressed to a voltage not exceeding 10 kV/in. When this unit is used in an oscilloscope it is screened by an aluminium cover that reduces the oscillator radiation whilst allowing sufficient ventilation.

7. Measurement of Rectifier Heater Operation

Measurement of the correct filament temperature of the high voltage rectifiers was made in a number of ways.

(*a*) The filament of a U37 valve was supplied from a known external variable supply and a high quality optical pyrometer with a close focus attachment was used to measure the temperature. Due to the small amount of light and the acute angle of observation this provided results which could not be consistently repeated, and was thus discarded.

(*b*) Accurately calibrated thermocouples and a suitable meter were used.[8] These gave good results when the thermocouple was isolated from the transformer e.h.t. windings, as when measuring valves when an output voltage of positive polarity is obtained at the filaments. However when measuring the filaments of valves supplying a negative output voltage the additional stray capacitance of the thermocouple and meter is placed across the transformer windings and so affects the operation of the transformer considerably. The method was discarded for this reason.

(*c*) Simple side-by-side comparison was made with one rectifier being supplied from a known source and the other being supplied from the e.h.t. unit. This gave an accuracy of between 5–10% and appears adequate.

8. Generator and Regulation Amplifier

For good efficiency it is desirable to operate the oscillator in Class C. The oscillator is connected in the Hartley configuration as this does not require critical circuitry.

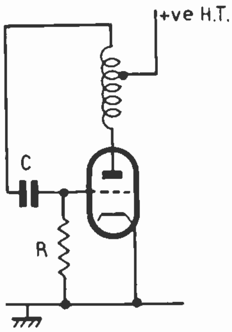


Fig. 4. Class C Hartley oscillator.

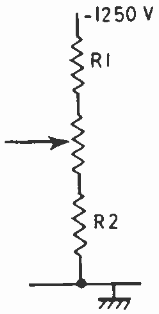


Fig. 5. Resistive attenuator sampling network.

The valve chosen was a beam tetrode with an anode dissipation of 12 watts (type 5763). In the design shown there is a considerable amount of allowable dissipation in hand on this valve, ensuring long life. By the use of the grid control of this valve, as described later, the screen grid may be held at a high potential and so the attainable g_m of the valve is kept at a high level.

The efficiency of the oscillator is

$$\frac{\text{power output of transformer}}{\text{power input to oscillator}} = 30\%$$

Whilst this efficiency is not very high, little difficulty will be encountered in providing the requisite h.t. supplies for normal low power application where the 5763 valve is used.

In a class C stage grid bias is usually obtained by a flow of grid current charging a time constant RC (Fig. 4). The grid acts as a rather poor diode, and on switching on the grid is at approximately zero volts, and a surge of current flows, some of which reaches the grid and drives it slightly negative. If the feedback is in the correct phase a voltage of the correct frequency is injected into the grid and drives it more

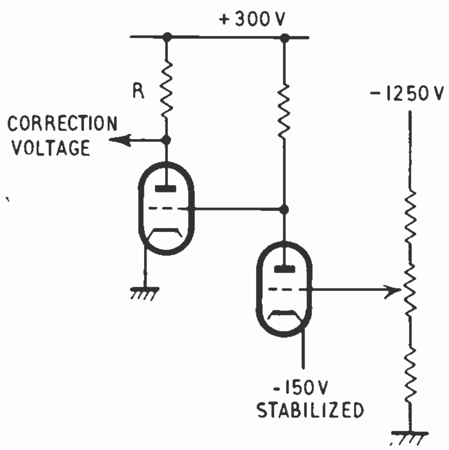


Fig. 6. Triode d.c. amplifier.

positive, the grid draws current and becomes slightly negative. It can be seen that as the first oscillation is only small the grid moves only slightly negative and as the oscillations increase the grid becomes more negative. The grid thus tends to stabilize oscillations and as there is virtually no bias on the valve on switching on, the g_m, and hence gain, is at a maximum and the circuit starts oscillations easily.

In the circuit shown it is desired to stabilize a negative supply of −1250 V. Two stages of amplification will be required in order to get the correcting voltage in the correct phase, i.e. when the stabilized line goes more negative the correcting voltage on the oscillator goes more negative and reduces the amplitude of oscillations.

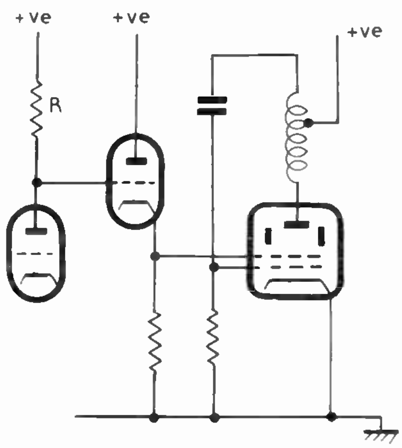


Fig. 7. Cathode follower feed for oscillator valve.

To sample the voltage which is to be stabilized a resistive attenuator is placed as shown in Fig. 5. The ratio R_1/R_2 is chosen to give a suitable output at the potentiometer for feeding into a d.c. coupled amplifier. In this instance $R_1/R_2 = 10$ and the sampled error voltage, which is one-tenth of the actual error voltage, is applied to the input of a triode d.c. amplifier (Fig. 6). The output of this triode is d.c. coupled into another triode and so the requisite polarity of voltage is obtained.

Having obtained a correcting voltage of suitable polarity there are several ways of applying this to the oscillator valve. The correcting voltage may be inserted into a cathode follower whose cathode feeds the screen grid of the oscillator valve (Fig. 7). Whilst the regulation is adequate there is the cost of an additional valve and the consequent increase in size.

If the anode resistor of the second d.c. amplifier is used as the screen feed resistor of the oscillator, Fig. 8, the cathode follower of Fig. 7 may be dispensed with, but R must be reduced in value in order

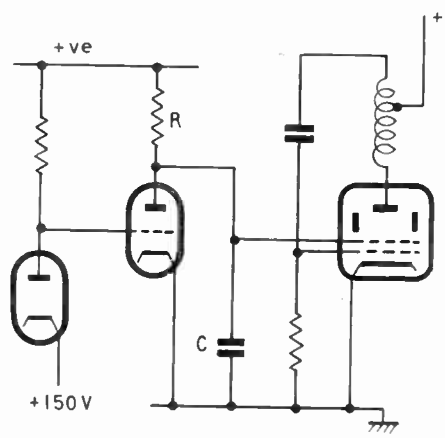


Fig. 8. Screen feed of oscillator.

to obtain sufficient current swing to vary the voltage on g_2 of the oscillator and also supply a suitable current to g_2 of the oscillator. It will also be heavily decoupled to earth via C, thus reducing the speed of response.

If, however, the correcting voltage is inserted into the control grid of the oscillator it is possible to obtain the higher gain and speed of response of Fig. 7 combined with the cheapness and simplicity of Fig. 8. The circuit of the grid controlled arrangement is shown in Fig. 9.

In the type of circuit shown in Fig. 10(*a*) the stabilized voltage cannot be corrected in a time faster than approximately $1/f$ seconds, where f is the frequency of the voltage source. This is shown in Fig. 10(*b*). If the stabilized voltage reduces at time A it will not be possible for the capacitor C to be recharged until time B, although during the time AB the oscillator amplitude is increased. If a faster response than $1/f$ is required a form of series stabilizer would be required.

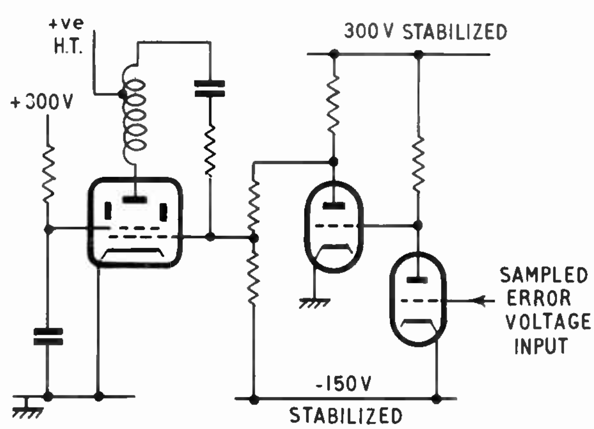


Fig. 9. Grid controlled oscillator.

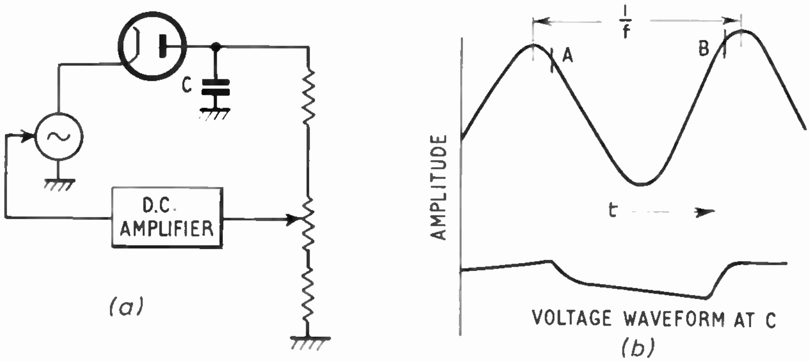


Fig. 10. Voltage stabilization.

A further improvement to the short term stability may be made by increasing the capacitance of C. It should be remembered however that the larger the capacitance of C the more dangerous the voltage becomes.

There is little difficulty in designing d.c. coupled amplifiers of suitable bandwidth for the regulator provided that suitable stable H.T. lines are available. Stabilization against supply variations is shown in Fig. 11. Reduction of heater voltage to both the oscillator and regulator simultaneously of up to approximately 30% made no apparent change in output voltage. The above reduction plus a reduction of 15% in the anode supply voltage of the oscillator valve lowered the maximum stabilized current from 800 μA to 700 μA.

In oscilloscope cathode-ray tubes the mean cathode current rarely exceeds 200 μA so there is more than adequate regulation in hand. It is essential that the −150 V line shown in the circuit (Fig. 12) is highly stable, as the sampled error voltage is compared against this.

9. Conclusions

The design of a compact economical stabilized e.h.t. generator has been shown. It is capable of supplying several differing high voltages and operates at frequencies greater than 40 kc/s, the one shown in Fig. 1 generating a stabilized supply of −1250 V, a d.c. coupled winding of −1520 V and a positive supply of 8·75 kV, and working at a frequency of 100 kc/s, and is suitable for a high quality c.r.o.

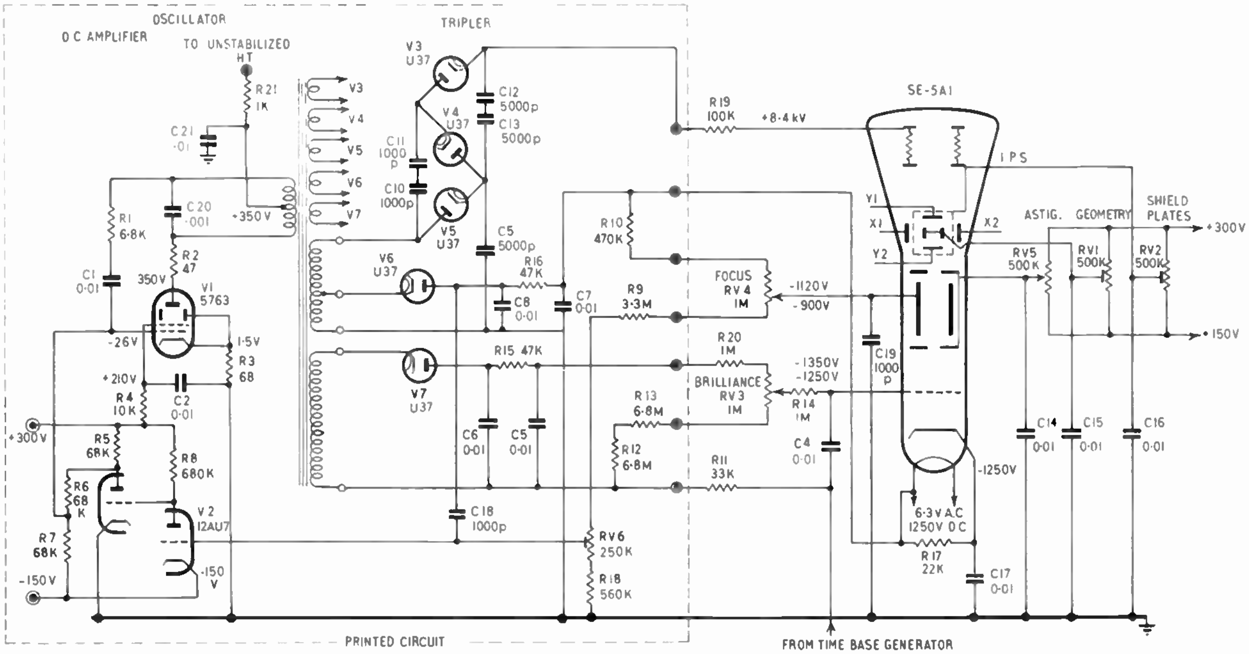


Fig. 12. Circuit of the power supply as used to feed an oscilloscope.

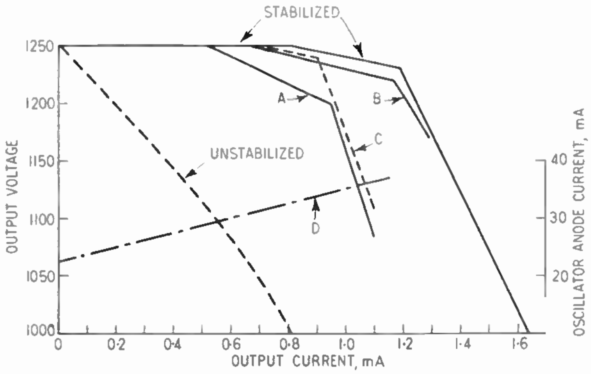


Fig. 11. Stabilization characteristics.

A Heater supplies reduced to 3·5 V a.c.

B Heater supplies reduced to 4 V a.c.

C Heater supplies reduced to 4·5 V a.c. and anode supply reduced from 350 V d.c. to 300 V d.c.

D Oscillator anode current at normal supply voltages.

The stabilized and unstabilized curves are shown with correct supply voltage.

Particular attention has been devoted to obtaining a transformer which is discharge free at the working voltage to ensure long life, and some factors affecting the reliability of transformer insulation have been shown. The considerably more stringent requirements for the insulation of a.c. potentials as compared to d.c. has been noted.

10. Acknowledgment

The author would like to thank Marconi Instruments Limited for permission to publish this paper.

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9. See p. 959 of ref. 4.
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APPLICANTS FOR ELECTION AND TRANSFER

As a result of its meeting on 22nd February the Membership Committee recommended to the Council the following elections and transfers.

In accordance with a resolution of Council, and in the absence of any objections, the election and transfer of the candidates to the class indicated will be confirmed fourteen days after the date of circulation of this list. Any objections or communications concerning these elections should be addressed to the General Secretary for submission to the Council.

Transfer from Associate Member to Member

BAILEY, Commander Arthur Joseph, R.N. *Portsmouth, Hampshire.*
McCRIRRICK, Thomas Bryce. *London, N.14.*

Direct Election to Associate Member

BROWN, Frederick Sydney. *East Bedfont, Middlesex.*
DANN, Paul Charles D., B.Sc.(Eng.). *Edgware, Middlesex.*
HILL, Denis Frank. *Edgware, Middlesex.*
HOOPER, Jack William. *North Harrow, Middlesex.*
*KILBURN, Lieutenant Commander George Charles J., R.N. *Guildford.*
LOCK, John Bainbridge. *Emsworth, Hampshire.*
WILLIAMS, John. *Liverpool.*
WOOLLEY, Captain Patrick John, R.E.M.E. *Reading, Berkshire.*

Transfer from Graduate to Associate Member

ARCISZEWSKI, Henryk, B.Sc.(Eng.). *Suresnes, France.*
BAGLEY, Aubrey Stephen. *Stanmore, Middlesex.*
BURGESS, David Albert. *Banstead, Surrey.*
CROOK, Maurice Howard. *Eastleigh, Hampshire.*
DAVIS, Eric Cambridge. *Sale, Cheshire.*
GEORGE, Prince Festus. *Freetown, Sierra Leone.*
HELLIWELL, Brian Stanley, M.A.(Cantab.). *Reading, Berkshire.*
HALL, Leo Frederick. *Nairobi, Kenya.*
HARKNETT, Maurice Richard. *Southsea, Hampshire.*
HOLLAND, John Phillip. *London, S.W.15.*
LEVY, Monty. *North Wembley, Middlesex.*
MADDOCK, Robert John. *Hedge End, Hampshire.*
POWRIE, Thomas Drew. *Stevenage, Hertfordshire.*
RENGARAYALU, Captain Subbarayalu, B.A. *Agra Cantt, India.*
TAYLOR, Albert Henry. *Liverpool.*
WALKER, David John. *Watford, Hertfordshire.*

Transfer from Student to Associate Member

MAY, Eric John Patrick, B.Eng., M.Eng. *Chorley Wood, Hertfordshire.*

Direct Election to Associate

ARCHER, Dennis Henry R., M.A.(Cantab.). *Harlow, Essex.*
BUCKINGHAM, Seymour Dennis. *East Croydon, Surrey.*
CROWE, Leonard Charles. *Sutton, Surrey.*
DODGINS, Robert Newlands. *South Shields, County Durham.*
*DU BERGER, Paul. *Quebec, Canada.*
SMITH, Lieutenant Commander Dennis Albert, R.N. *Petersfield, Hampshire.*
SUGGARS, Ronald Edward. *Windsor, Berkshire.*

Direct Election to Graduate

AKBAR, Flight Lieutenant Syed Mohamad, M.Sc., Dip.El. *Risalpur, Pakistan.*
BANSTEAD, Edward James. *Wickford, Essex.*
BASSETT, Richard. *Cheltenham, Gloucestershire.*
BOOTY, Dennis Sydney. *London, S.W.6.*
BRAY, Alan John. *Carshalton, Surrey.*
BRITTEN, John Henry. *Bath, Somerset.*
BURGESS, Brian Douglas. *Chelmsford, Essex.*
CARPENTER, John Alexander. *Beckenham, Kent.*
COLE, Christopher Brian. *Plymouth, Devon.*
CROOKS, Captain Peter Victor, B.Sc., R.E.M.E. *Reading, Berkshire.*
CRYER, Terence Frederick. *Harrogate, Yorkshire.*
EDWARDS, Kenneth Percival. *London, N.W.9.*
FARTHING, Hugh. *Cardiff, Glamorgan.*
FILKINS, Roy. *Welling, Kent.*
GEALL, Anthony Reginal. *London, E.6.*
GROVES, Arthur Derek George. *Knebworth, Hertfordshire.*

Direct Election to Graduate (*cont.*)

HARRISON, William Matthew. *Nottingham.*
HAWKINS, Victor Stanley Wilfred. *Cheltenham, Gloucestershire.*
HICKMAN, Thomas Roy. *Dudley, Worcestershire.*
HOLDEN, Brian. *Reading, Berkshire.*
HUGHES, Robin Morris. *Uxbridge, Middlesex.*
JAMES, Albert Henry. *Worcester.*
JAMES, Anthony Trevor. *Reading, Berkshire.*
JEFFERYS, Peter Leonard. *Watford, Hertfordshire.*
KINGSWOOD, Leslie Douglas. *Horley, Surrey.*
LINDSLEY, David Maurice. *Bognor Regis, Sussex.*
LOCK, Roger David. *Edgware, Middlesex.*
LUCY, George Thomas. *Reading, Berkshire.*
MACKAY, Robin Graham. *Stevenage, Hertfordshire.*
McLEAN, Roy Alfred. *London, E.4.*
McMILLAN, Alan James. *Nottingham.*
MARZOLINI, Remo G. A. *London, W.C.1.*
MATTHEWS, Hector McDonald John. *Romford, Essex.*
NARANG, Ravinder Pal S., M.Sc. *Slough, Buckinghamshire.*
NELLIS, Juri. *Chelmsford, Essex.*
NOBLE, Peter John Wellesley. *Reading, Berkshire.*
OLDLAND, Brian William. *Bracknell, Berkshire.*
ONGLEY, Gordon Clive. *Bognor Regis, Sussex.*
PEARCE, Thomas Leslie. *Basingstoke, Hampshire.*
PEMBERTON, Robert. *Hatfield, Hertfordshire.*
PILCHER, Berry. *Northfleet, Kent.*
PREES, Michael Ronald. *Cambridge.*
RAAB, Anthony Rowland, B.Sc. *Gurnard, I.O.W.*
REYNOLDS, Maurice. *Hitchin, Hertfordshire.*
ROBERTS, Brian Gordon. *Rayleigh, Essex.*
SMIRK, Fred. *Cheltenham, Gloucestershire.*
STANNERS, Norman Dennis. *High Wycombe, Buckinghamshire.*
TAYLOR, Derek Henry. *Bury St. Edmunds, Suffolk.*
TAYLOR, Raymond Jack. *Reading, Berkshire.*
WEBB, David John. *Cheltenham, Gloucestershire.*
WHITE, Christopher Robina. *Great Baddow, Essex.*
WILES, James Robert. *Romford, Essex.*

Transfer from Student to Graduate

AHERN, Brian Henry. *Portsmouth, Hampshire.*
ALDERSLADE, David John, B.Sc.(Eng.). *London, N.10.*
CALDICOTT, Jack Richard. *Coventry.*
CHILDE, Percy. *Hong Kong.*
CHOY WAI MAN. *Hong Kong.*
COLES, Ronald David Charles. *Nuneaton, Warwickshire.*
CRAWFORD, Brian Thomas. *Stevenage, Hertfordshire.*
DARUKHANAWALA, Yezdi Dinshaw. *Bombay.*
GARRIOCH, James Traill Moodie. *Sanday, Orkney.*
GOUGH, Kenneth Ernest. *Epping, Essex.*
HALL, Ernest Charles. *Basildon, Essex.*
HARRISON, William. *Oldham, Lancashire.*
HARWOOD, Anthony James. *Reading, Berkshire.*
KEANE, John Richard, B.Sc. *Woodley, Berkshire.*
McCONNELL, John Frank. *Cambridge.*
McKISSOCK, James Barr. *Catterick Camp, Yorkshire.*
MACLEAN, Alexander Murdo. *London, W.5.*
MARCHANT, Raymond Jack. *Southampton, Hampshire.*
MEHTA, T. Raj. *New Delhi, India.*
NICHOLAS, Anthony Michael. *Plymouth, Devon.*
PALMER, Donald Valentine. *Seven Islands, Canada.*
PRABHAKARA RAO, Captain Gadiyar. *Mhow, India.*
RUSHWORTH, Alan. *Basingstoke, Hampshire.*
SCURRAH, Robert Eric. *Cleckheaton, Yorkshire.*
SEENEY, Gordon William. *B.F.P.O. 10.*
TIERNEY, James Clive Cameron. *Cambridge.*

STUDENTSHIP REGISTRATIONS

The following students were registered at the January meeting of the Committee. The names of a further 85 students registered at the February meeting will be published later.

ADEFEMI, Stephen A., B.Sc. *London, S.E.27.*
*BHAT, Manohar Kashinath. *Bombay.*
BHOJANI, Natwarlal P., B.E. *London, W.10.*
BISHOP, Gilbert D. *High Wycombe, Bucks.*
CHAPMAN, Ronald A. *Thurso, Caithness.*
CHONG, Randolph, T. Dip.El. *Singapore.*
*CHOPRA, Om Prakash. *Delhi.*
COPPING, John Richard. *Portsmouth.*
DAY, Geoffrey George. *Croydon, Surrey.*
FRIMPONG, Kwadwo Darbo. *London, N.16.*
GALEA, Thomas T. *Malta, G.C.*
GARVIN, Michael John. *Cambridge.*
GORDON, Ian Francis. *London, E.C.3.*
GUNASEKERA, Rajapaksepathirage Hector W. *Ceylon.*
HARVEY, Frederick, J. *Greenford, Middlesex.*
KAM SENG LAU. *Singapore.*
KELLY, Gordon K. *Southend-on-Sea, Essex.*
KISHON, Yehezkiel, M.D. *Ramat Gan, Israel.*
LEACH, Peter A. G. *Arabian Gulf.*
MAHIDHAR, Ajit Kumar, B.Sc. *Bombay.*
MAIDOH, Gabriel Blackie. *Lagos.*
NEWMAN, David W. *Basingstoke, Hampshire.*
ONIANWA, Agboso. *Southampton, Hampshire.*
*PANT, H. C., B.Sc. *Dehra Dun, India.*
PHULL, Joginder Singh, B.A. *Nairobi.*
*RADHADKRISHNAN NAIR. *Kerala, S. India.*
RENDLE, Gerald. *Plymouth, Devon.*
ROSSI, Francis. *Malta, G.C.*
*RUPRAI, Balwant, S. *Secunderabad, India.*
SAX, Nicolaas. *Suva, Fiji Islands.*
*SCOTT, Walter. *London, N.6.*
SELVARAJ, John David. *Madras.*
SIDDIQI, Wasi Ahmed, B.Sc. *Karachi.*
SOUTHALL, Geoffrey R. *Kidderminster.*
WHITNEY, John Gerrard. *London, S.W.20.*
WILLCOCKS, Gary Henry. *London, S.W.6.*

* Reinstatements.

U.D.C. 629.19: 621.396.005.1

An Economical Satellite Launching Technique for Conducting Radio Research in Space

By

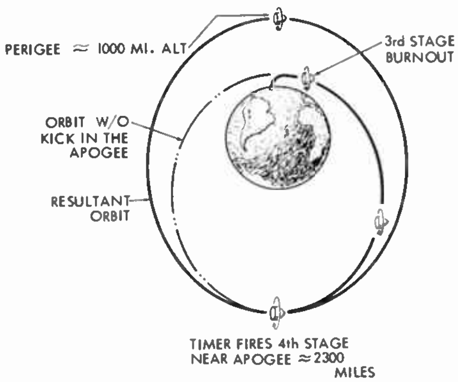
JOHN D. NICOLAIDES,
M.S.E.†

Presented at the Convention on "Radio Techniques and Space Research" in Oxford on 5th–8th July 1961.

Summary: A simple and inexpensive launching technique for economical radio research is described which employs an aircraft and rocket combination to place a 20 lb satellite into orbit. The paper reviews the progress to date and indicates performance potential.

1. Introduction

The gigantic advance in astronautics is now dominated by bigger boosters directed towards larger payloads for deeper space probes. While this effort may be worthwhile in the long run, it is certainly true that there now exists a critical need for a very inexpensive technique for launching small satellites of weights ranging from 10 lb to 50 lb. Such a launching technique would have general world-wide utility since each laboratory could carry out its own programme of scientific research and space applications, independently or jointly as desired.



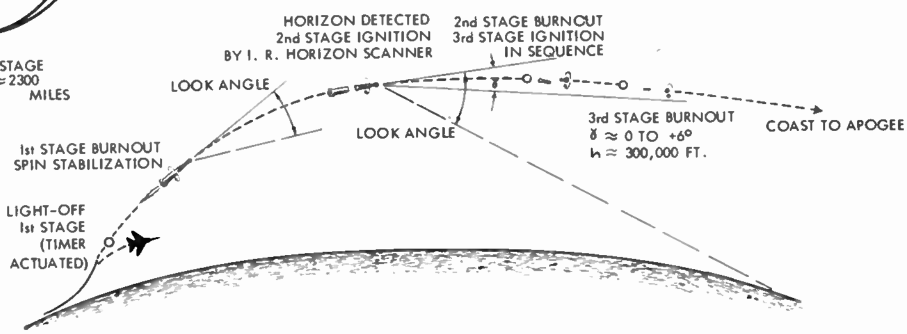


Fig. 1. The technique for launching satellites from aircraft.

The purpose of this paper is to indicate one such inexpensive small satellite launching technique and also to speculate on a few possible payloads. It is recognized that there are many payloads of interest and value to radio engineers and that many more will evolve in the future.

2. Aircraft Launching of Satellites

The technique which is suggested here[1] proposes the use of aircraft as a launching platform and employs a special 4-stage rocket vehicle. The aircraft carries the rocket vehicle to high altitude above the dense portions of the atmosphere. The aircraft then performs a steep climb manoeuvre and releases the rocket vehicle at the proper initial angle and with considerable initial velocity. After the rocket vehicle is clear of the aircraft, its first stage ignites and the vehicle is carried to high altitude. Due to fin stabilization, the vehicle will trail to a horizontal flight position. At this point, the spin-stabilized second stage is separated and its rocket ignited. After burning the second stage drops away and the third stage is ignited. Thus, the large velocity so added permits the payload to travel around the earth in an elliptical orbit with a low perigee altitude equal to the altitude of the termination of thrust of the third stage. Since this perigee altitude is low and will therefore cause the satellite to spiral in and re-enter the atmosphere, additional velocity must be added at apogee (Fig. 1). Since the satellite is spin stabilized, and therefore maintains the same angular orientation in space at all times, a fourth stage motor, originally oriented backwards, is ignited, thus adding velocity at apogee to achieve a circular orbit at that high altitude.

† Technical Director for Astronautics, U.S. Bureau of Naval Weapons; now Director of Program Review and Resources Management, Office of Space Sciences, National Aeronautics and Space Administration, Washington 25, D.C., U.S.A.

It is important to note that all the stages of the vehicle contain simple solid propellant motors, which can be stored for long periods of time, and have a high degree of reliability. Since the vehicle is primarily a ballistic vehicle, no moving parts are involved in its design and flight programming. A typical vehicle, shown in Fig. 2, has an approximate weight of only 3000 lb. It can be launched from a standard aircraft and it has the capability of placing a 10 lb payload in a 1000 mile polar orbit, or a 25 lb payload in a 300 mile orbit of 30 degrees inclination. The cost of the vehicle is approximately 50,000 dollars in small lot production.

Six launchings were carried out in the autumn of 1958. The design of these original vehicles was quite different from the later proposals in that a cluster of four motors was used as the first and second stages. It cannot be stated with absolute assurance that any of these vehicles placed a satellite in orbit. However, the data so obtained did clearly indicate the basic feasibility of such an approach, and the knowledge gained led to an optimum redesign of the vehicle.[1] The new vehicle has had two full-scale tests employing, however, only a live first stage. The results of these tests clearly indicated that the basic design is sound and that additional shots can now be programmed with live upper stages. This vehicle also has the capability of being used for probes to high altitudes, or long ranges.

The performance of the system can, of course, be improved, if desired, by using higher performance aircraft. Also, the development of larger stages readily lends itself to further vehicle improvement which could clearly launch payloads of from 50 to 100 lb in useful orbits. Thus, the air-launch system has considerable growth potential leading ultimately to a specially designed aircraft as was recently suggested by Dr. Theodore Van Karman and by certain early U.S.S.R. films.

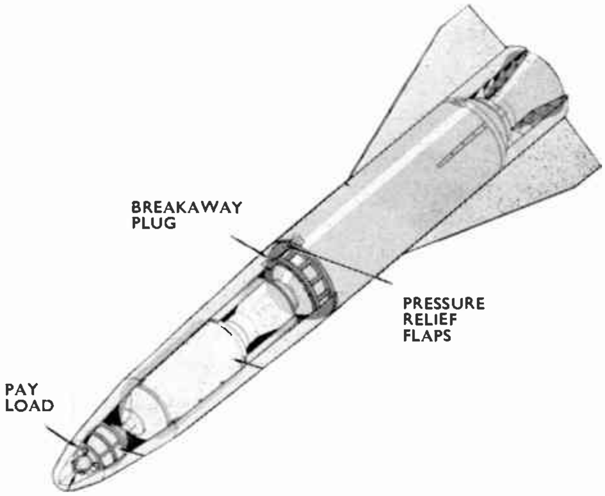


Fig. 2. The 4-stage air-launched vehicle.

3. Some Conceivable Small Satellites

3.1. *Space Observatory*

This is a manoeuvrable telescope which can be launched as a vertical probe by the air-launched vehicle for the purpose of making various measurements in outer space of radiations coming from the sun, the firmament, and the earth. The satellite shown in Fig. 3 is spin stabilized so as to maintain its attitude in space and employs nutation dampers to remove any wobbling motion. Also, its orientation in space is controlled by small pulsating motors, which are capable of both precessing this spinning satellite as well as translating it.

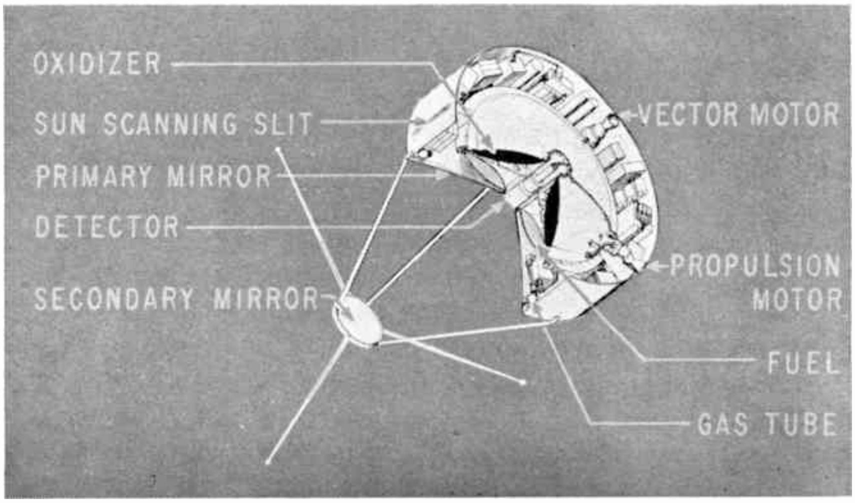


Fig. 3. The manoeuvrable telescope satellite.

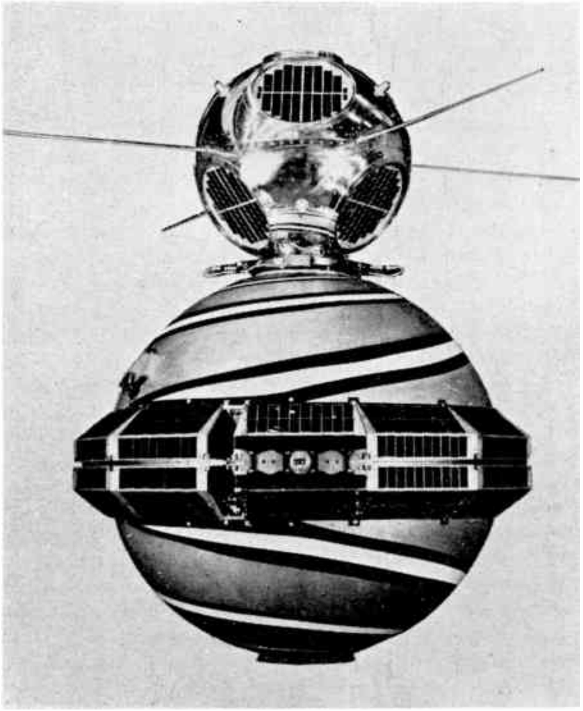


Fig. 4. The *Transit IV-A* satellite.

3.2. *Balloon Satellite*

A 20-ft balloon satellite has been proposed for launching by the techniques described. Such a satellite would be useful in experimental communications work, in studying atmospheric drag, and in determining the effects on satellite motions of solar pressure. There have also been suggestions that small payloads attached to such a balloon satellite could be recovered from orbit since its ballistic coefficient enables ready passage through the critical re-entry phase.

3.3. *Ionospheric Beacon Satellite*

An ionospheric beacon satellite is also possible based on the concepts evolved from Project *Transit*[2] (Fig. 4). Such a small satellite of 25 lb could transmit six coherent frequencies derived from a single highly stable oscillator. A research laboratory, observing these six frequencies and using the formula given below, could make a determination of the unknown constants as a function of time and therefore undertake a scientific study and a time-dependent mapping of the ionosphere

$$f_R - f_S + \frac{f_S}{c}\dot{\rho}(t) = \frac{A(t)}{f_R} + \frac{B(t)}{f_R^2} + \frac{C(t)}{f_R^3} + \frac{D(t)}{f_R^4} + \ldots$$

where f_R = frequency received by ground station

f_S = frequency actually transmitted by satellite

c = propagation velocity of electronic wave

ρ = absolute instantaneous distance between satellite and ground station

A, B, ... = ionospheric parameters.

3.4. *Rescue Satellite*

SARUS[3] is a proposed small (20 lb) rescue satellite which would receive signals from downed aircraft or from ships in distress, store this information and then transmit it back to a central station for analysis and determination of the location of the downed aircraft or distressed ship.

3.5. *Infra-Red Scanner Satellite*[4]

Cloud cover pictures could be taken by the infra-red scanner satellite (Fig. 5) which has been launched by other techniques. Weighing 7 lb, this satellite is spin stabilized and contains a nutation damper to eliminate all wobbling motion. The scanning technique, combined with continuous transmission of received signal, permits the design of a simple and reliable satellite. Various filters permit concentrated attention on special bands of interest.

4. Concluding Remarks

It is recognized that these have been simple and obvious examples of a few possible satellites which could be launched by this vehicle. It is certainly recognized that there are many other possibilities which radio engineers would be interested in pursuing. The big limitation which exists today is the un-

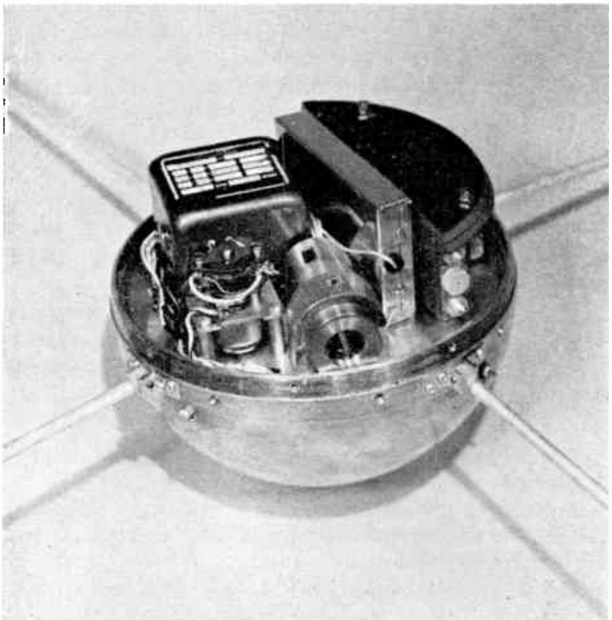


Fig. 5. The infra-red scanner satellite.

availability of such a simple, inexpensive, mobile and flexible launching vehicle. The system set forth here offers the possibility for any laboratory in any country to carry out those space research and application programmes in which it is interested. Further, such a technique offers the opportunity for each of those countries to improve on the performance of the vehicle, either by using higher performance aircraft which are available to them, or by simple redesign of some of the stages of the vehicle. It is clear that such work could lead towards a greatly improved system capable of launching satellites certainly of 50–100 lb weight and even extended to much greater systems, depending upon the design of the unique aircraft and launching vehicle involved.

5. References

1. G. F. Cleary, "Preliminary Design of a Simple Low-Cost Satellite-Vehicle Launched from an Aircraft", Naval Ordnance Test Station, August 1959.
2. J. D. Nicolaides, "Project TRANSIT", *Aerospace Engineering*, February 1961.
3. P. J. Klass, "N.A.S.A. considers search-rescue satellite", *Aviation Week*, June 29, 1959.
4. G. McCarty, "N.O.T.S. IR Scanner", Naval Ordnance Test Station, 1959.

Manuscript received by the Institution on 2nd June 1961. (Paper No. 711.)

POINTS FROM THE DISCUSSION

Lt.-Col. J. A. D. McEwen (*Member*): Guidance requirements during the injection phase have not been mentioned in detail. Do not small errors which appear possible cause large errors in apogee?

The author (*in reply*): It is true that small errors introduced at perigee can cause large errors at apogee. Fortunately these errors can be computed and requirements for the conditions at perigee can be established. Our analyses indicate that we can maintain these errors small enough as not to present a serious problem. Of course, the real proof is in actual flight testing.

Mr. R. Rowland: Would the author reconcile spin scanning with spin stabilization—it appears that perpendicular scanning is confined to two small regions.

The author (*in reply*): Spin stabilization is just the old ballistic art of gunnery. In space, however, the forces are very small and thus a spinning satellite is, in fact, a perfect frictionless gyroscope. If an electronic eye looks out of the side of this spin stabilized satellite it sees successive spots on the ground as it scans from horizon to horizon. The forward motion of the satellite allows successive lines to be scanned, in much the same way as a television picture is presented.

It is true that useful pictures may be confined to two regions of the earth. The size of these regions, of course, depends upon your particular design of instrumentation and the altitude of the satellite.

Mr. K. E. V. Willis: In the recently announced "triple" satellite launching in the U.S.A., it was stated that one of the satellites used a "nuclear battery". Was a nuclear battery used or was some non-fissile material in conjunction with thermo-electric elements employed?

The author (*in reply*): The *Transit IV-A* satellite contains a nuclear auxiliary power source—a small, light-weight, radioisotopic-fuelled thermoelectric generator developed for the Atomic Energy Commission under contract with the Martin Company. The generator, about the size and shape of a grapefruit, will provide a small amount of direct electrical current to instrumentation and two of the four transmitters in the satellite. Its fuel is theoretically sufficient to provide continuous power for many years.

Mr. H. H. J. Gorham: The author states that by using higher performance aircraft the system performance can be improved. How much improvement does he envisage by this means? Secondly, how large would his larger rocket system be and would he define his "useful orbits" for 100 lb loads?

The author (*in reply*): The figures given in the paper apply to the F4D aircraft. If the F4H is used, preliminary estimates indicate that a 20 lb payload may be placed in a 1000 mile polar orbit and 40 lb payload may be placed in 500 mile equatorial orbit.

By employing higher performance motors and increasing the vehicle weight by a factor of almost 3 but still employing the F4H, a payload of approximately 75 lb may be placed in a 1000 mile polar orbit and a payload of approximately 125 lb may be placed in a 500 mile equatorial orbit. Of course, there are much larger aircraft with greater carrying capability which could launch much larger payloads into higher orbits. However, it is my considered judgment that important satellites in the areas of navigation, weather, communication, and scientific research can be constructed at the 100 lb weight level and that this aircraft launching technique deserves some consideration.

Mr. C. M. Cade (*Member*): Mention was made by Mr. Nicolaides of v.l.f. transmissions to underwater receivers; does not this mean penetration of ionosphere by waves below the nominal critical frequency? Surely under these conditions reception is limited to a very small area of earth's surface?

The author (*in reply*): Very definitely this means penetration of the ionosphere by a wave below the nominal critical frequency. Specifically, we transmitted at 18 kc/s. Also, this LOFTI satellite designed and developed by the Naval Research Laboratory has already demonstrated the feasibility of broad area coverage at ranges as great as 10 000 miles.

This space research work offers great promise in extending our knowledge of the ionosphere and in leading to new space applications. This is the kind of discovery which small satellites can readily make.

U.D.C. 621.383.8: 621.391.883.2: 527

Signal-to-Noise Power Ratio available from Photomultipliers used as Star Detectors in Star Tracking Systems. A method of assessment

By

Squadron Leader

D. S. J. CHAPMAN

(Associate Member)†

Summary: In space navigation, or attitude control of satellites, where guidance is effected by reference to a set of axes defined by a stabilized platform, continuous monitoring of the reference axes will be required in order to compensate for the inevitable long-term drift of the stabilizing gyros. Changes in the directions of distant "fixed" stars provide a convenient monitoring reference; in order to use such a monitoring reference, two or more star tracking systems must be employed. The concept of a star tracking system as (*a*) a direction measuring device demanding, for satisfactory operation, a minimum signal/noise ratio, and (*b*) a star detecting device providing a signal/noise ratio which is dependent upon certain system parameters, is briefly stated. A method of deriving an asymptotic approximation to the available signal/noise ratio as a function of stellar visual magnitude is described; in deriving this asymptotic function, the relationships between a number of system parameters are clarified and given graphical significance.

List of Symbols

a	photomultiplier noise enhancement factor.
A	angular area of the field of view afforded by a focal plane aperture—square degrees.
B	bandwidth of photomultiplier and associated amplifier—cycles/second.
e	electron charge—$1{\cdot}6 \times 10^{-19}$ coulombs.
F	photomultiplier collection efficiency.
G	photomultiplier current multiplication factor.
i_{de}	equivalent dark current including effect of thermal noise—amps.
i_d	photomultiplier cathode dark current—amps.
i_s	photomultiplier cathode signal current—amps.
i_t	total photomultiplier cathode current—amps.
I	sky background luminance—lumens per square metre of entrance pupil area per steradian (M.K.S. units).
k	Boltzmann's constant—$1{\cdot}372 \times 10^{-16}$ ergs/degree.
K, K'	numerical constants.
M	equivalent visual magnitude of sky background illuminance.
m_v	stellar visual magnitude.
n	expectation, or average number of stars in a given angular area.
p	effective optical entrance pupil area—square inches.
P	probability.
R	photomultiplier load resistance—ohms.
t	time in seconds.
T	temperature—degrees K.
δ, Δ	angular length—arc seconds.
ϕ_B	sky background illuminance—lumens per unit area of entrance pupil.
ϕ_S	star illuminance—lumens per unit area of entrance pupil.
σ	photomultiplier cathode sensitivity—amps per lumen.

† Air Ministry, Whitehall, London, S.W.1.

1. Introduction

Navigation may be defined as a method of determining the relationship in space between the navigator and his destination. In order to describe uniquely such a relationship, the navigator seeks range and direction relative to some set of axes through some point of origin. In earthbound navigation, the set of axes defining "up-down", "North-South", and "East-West" through the position of the navigator can be provided by instruments operated by gravity and the earth's magnetic field. In free space these operating forces are not available for use; however, a set of reference axes can be provided in the form of a gyro-stabilized platform.

Despite the very high degree of accuracy which can be achieved in the manufacture of stabilizing gyros, for long-term navigation in space, or for the attitude control of satellites, some means of monitoring the inevitable drift of the gyros is necessary, and a convenient monitoring reference is available if one can observe any changes in the directions of distant "fixed" stars. Observation of the direction of a single star yields two-dimensional information (i.e. an observed error can be resolved into errors about two axes normal to the line of sight, but gives no information regarding any error about the line-of-sight axis)

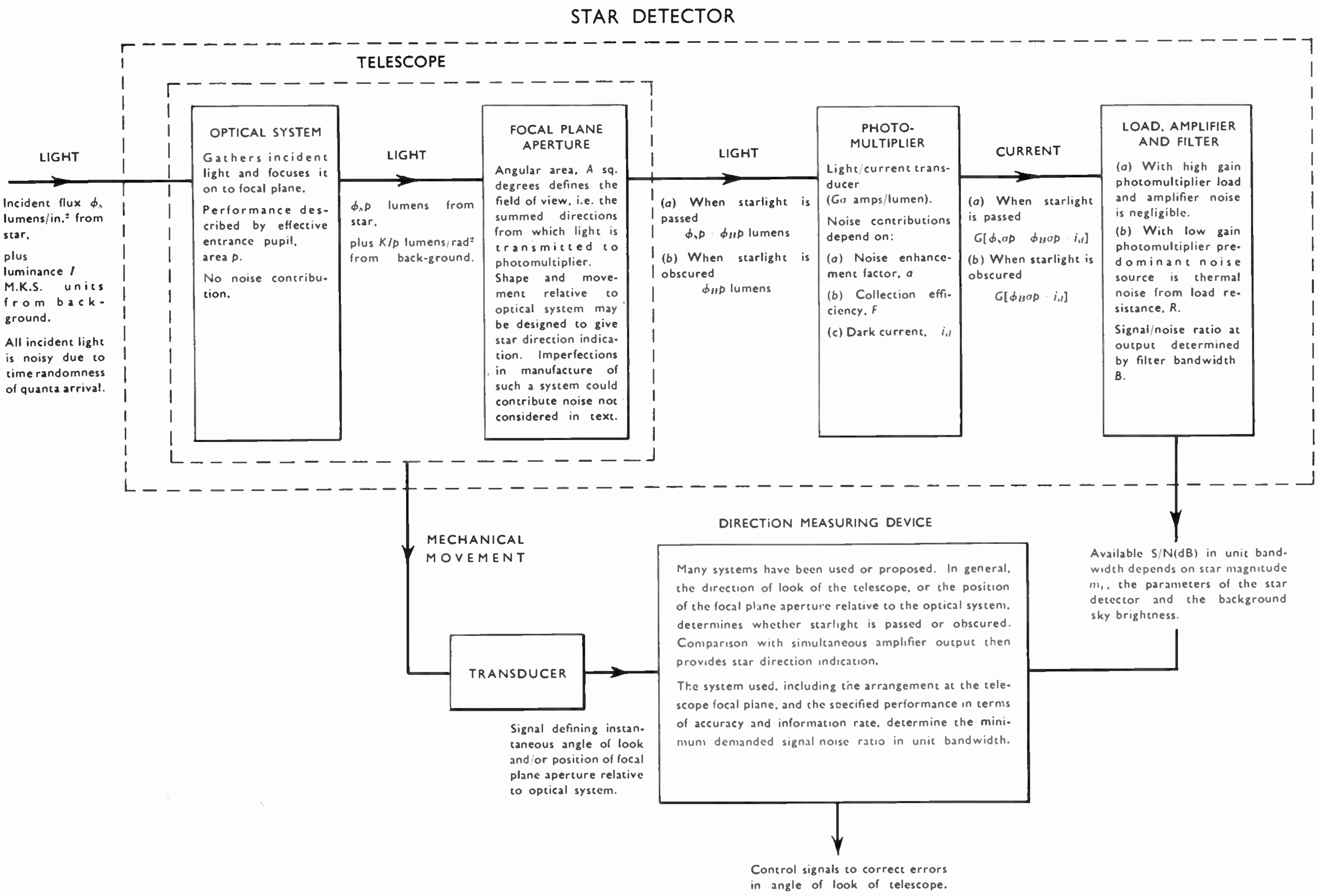


Fig. 1. Essential components of a star tracking system.

and therefore for each record of platform attitude at least two fixed stars must be observed. The function of a star tracking system is to observe and measure the two-dimensional error in the direction of a star relative to the expected direction.

2. Demanded Signal-to-Noise Ratio as a Function of Performance Specification

In any particular application a star tracking system will be required to indicate a direction error to some specified accuracy within some specified time; the specification of accuracy and information rate will depend upon the application and the rate of gyro drift. When (as is the case when a photomultiplier tube is used) the detecting device has only one photosensitive surface, the means of measuring direction must include some form of scanning. In this case the specification can be expressed as a minimum demanded signal/noise power ratio in unit bandwidth. A simple example will serve to clarify this.

Suppose the accuracy of the stabilized platform is such that the monitoring star is known to be within an area of the celestial sphere which can be scanned with a line δ arc seconds wide and Δ arc seconds long, and that the direction of the star must be measured to an accuracy of δ arc seconds every t seconds of time. The scan speed must be no faster than will allow a signal pulse to rise to some detectable level above noise in the time taken to traverse δ arc seconds, i.e. in $t.\delta/\Delta$ seconds of time. Since the pulse rise time is inversely proportional to the bandwidth of the detector and amplifier, the specification of accuracy and information rate sets a limit to the narrowness of the system bandwidth, and hence to the amount of noise limitation which can be effected by filtering. In the case of this example the minimum demanded signal/noise ratio in unit bandwidth could be expressed as

$$K+10\log\left(\frac{\Delta}{t\delta}\right)\text{ decibels}$$

i.e. as a function of the performance specification.

There is, of course, a very large variety of possible systems of scanning and signal processing which have been used or proposed. In general, however, for a given system in a particular application, a figure can be calculated for the minimum demanded signal/noise ratio in unit bandwidth which will provide satisfactory performance.

3. Factors Affecting the Available Signal-to-Noise Ratio

The task of the designer, having been given a specification for performance, is to postulate a system in which the signal/noise ratio demanded by the direction measuring device can be met by the star detector.

The essential features of the star detector are illustrated at Fig. 1, and may be summarized as follows:

(a) An optical system to gather and focus the incident light. Its performance can be expressed by a single parameter p, the effective entrance pupil area.

(b) An aperture at the focal plane of the optical system restricting the instantaneous field of view. The field of view is conveniently described by the area A, in angular measure, of the focal plane aperture.

(c) A photomultiplier tube to detect the light passing the focal plane aperture. Its essential characteristics are:

  (i) Cathode sensitivity.
  (ii) Current multiplication factor.
  (iii) Cathode dark current.
  (iv) Collection efficiency.
  (v) Noise enhancement factor.

(d) The photomultiplier collector load resistance and associated amplifier.

Each of these features has an influence upon the signal/noise ratio which is available from a specific star.

Since it may be necessary to use any one of a large number of stars of different brightness it is convenient to consider the available signal/noise ratio as a function of stellar visual magnitude. It should be noted here that stars are normally catalogued according to their visual magnitude, m_v, which expresses the light flux density perceived by the human eye. Because of this it is convenient to express the sensitivity of the photomultiplier cathode in amperes per lumen.

In any detailed analysis it should be borne in mind that the use of the lumen and "visual magnitude" assumes "black body" stars of specific temperature, or a photomultiplier cathode having a spectral sensitivity exactly matching that of the human eye. Since, in general, neither of these conditions apply, corrections are needed for any particular cathode material, and for each particular star according to its colour temperature. Photomultipliers with antimony-caesium cathodes are commonly used, and in this case, on stars of average colour temperature, the correction required is small; it has therefore been disregarded in the present context which is concerned primarily with assessment to a first-order accuracy.

4. The "Ideal" Signal/Noise Function

A photomultiplier cathode, illuminated by a star, emits a cathode signal current consisting of a time-random flow of electrons each of charge e. For such a time-random flow the well-known "shot-noise" equation is valid, so that for a signal current of mean value $\bar{i}_s$ the mean square noise current in unit bandwidth is $2e\bar{\imath}_s$. Hence the signal/noise power ratio in unit bandwidth will be given by

$$S/N(\text{dB}) = 10 \log \left[\frac{\bar{\imath}_s^2}{2e\bar{\imath}_s}\right] = 10 \log \frac{\bar{\imath}_s}{2e} \quad \ldots\ldots(1)$$

If the optical system has an effective entrance pupil area p in^2, and the sensitivity of the photomultiplier cathode is σ amperes per lumen, an incident signal flux density of ϕ_S lumens/in^2 will yield a signal current $\bar{\imath}_s$ equal to $\phi_S\sigma p$. Hence

$$S/N(\text{dB}) = 10 \log \left[\frac{\sigma p}{2e}\right] + 10 \log \phi_S \quad \text{in unit bandwidth} \quad \ldots\ldots(2)$$

One standard candle at a distance of 1000 metres (an illuminance of 10^{-6} lumens/m^2 or $6{\cdot}452 \times 10^{-10}$ lumens/in^2) produces a visual magnitude of $+0{\cdot}8$. The scale of visual magnitude is such that its numerical value is decreased by 1 when the illuminance is increased by a factor of 2·5. Hence, from a source of visual magnitude m_v, the illuminance, ϕ_S, is given by

$$\phi_S = \frac{6{\cdot}452 \times 10^{-10}}{2{\cdot}5^{(m_v - 0{\cdot}8)}} \text{ lumens/in}^2 \quad \ldots\ldots(3)$$

Whence

$$10 \log \phi_S = -88{\cdot}703 - 4m_v \quad \ldots\ldots(4)$$

The signal/noise ratio can, from eqns. (2) and (4) be written as a function of the signal star visual magnitude. Thus

$$S/N(\text{dB}) = 96{\cdot}25 + 10 \log(\sigma p) - 4m_v \quad \text{in unit bandwidth} \quad \ldots\ldots(5)$$

This function accounts only for shot noise on the signal illuminance itself. In practice photomultipliers contribute additional noise[1] for two reasons.

(a) All the signal current will contribute to noise at the output since the time-randomness of the individual electrons leaving the cathode is repeated by the time-randomness of the current pulses from the collector; however, all the cathode signal current may not appear in magnified form as output signal current. The device has a collection efficiency F, which is less than unity and typically 0·9.

(b) The photomultiplier gain G suffers random variation, so that for a mean square signal current $G^2\bar{\imath}^2{}_s$, the mean square noise current is $2e(Ga)^2\bar{\imath}_s$ where a is the noise enhancement factor. a is quoted[1] typically as 1·15.

The additional photomultiplier noise degrades the signal/noise function by $10 \log \left[\frac{F}{a^2}\right]$ dB, which is typically less than 2 dB. An "ideal" signal/noise function could therefore be described by

$$S/N(\text{dB}) = 94{\cdot}25 + 10 \log(\sigma p) - 4m_v \quad \ldots\ldots(6)$$

which is a straight line of slope -4 dB per magnitude (Fig. 2).

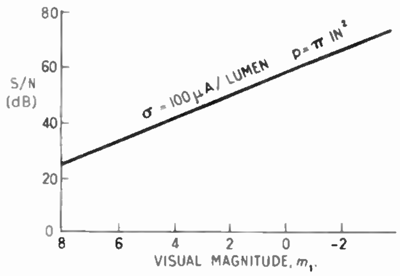


Fig. 2. "Ideal" signal/noise function.

5. Asymptotic Approximation to the Operational Function

Under operating conditions further contributions to noise will arise from:

(a) Shot noise on the dark current.

(b) Shot noise on current due to sky background luminance.

(c) Thermal noise current due to having to work into a load having real finite resistance, R.

(d) Scintillation due to atmospheric conditions. This factor is included for completeness, but it will contribute noise only in the special case of the star tracker operating within the atmosphere. In such a case its importance may be paramount, but its consideration is not within the scope of this paper.

The total mean square shot noise current at the output will be $2e(Ga)^2B\bar{\imath}_t$ where $\bar{\imath}_t$ is the mean total cathode current; also the thermal noise current at an ambient temperature T will be $4kTB/R$. Hence, disregarding scintillation noise, the total mean square noise current at the output will be

$$2e\left[(Ga)^2\bar{\imath}_t + \frac{2kT}{eR}\right] \text{ in unit bandwidth}$$

$$= 2e\left[(Ga)^2\bar{\imath}_t + \frac{1}{20R}\right] \begin{array}{l}\text{in unit bandwidth at} \\ \text{a normal temperature}\end{array} \ldots\ldots(7)$$

The mean square signal current at the output is $FG^2\phi_S^2\sigma^2p^2$ so we can write

$$S/N(\text{dB}) = 10 \log \left[\frac{F}{2ea^2} \cdot \frac{\phi_S^2\sigma^2p^2}{\bar{\imath}_t + \dfrac{1}{20Ra^2G^2}}\right] \quad \text{in unit bandwidth} \quad \ldots\ldots(8)$$

$\bar{\imath}_t$ consists of mean signal current, mean dark current, and mean current due to sky background luminance, and can be written

$$\phi_S\sigma p + i_d + \phi_B\sigma p$$

Now if

$$\phi_S\sigma p \gg \phi_B\sigma p + i_d + \frac{1}{20Ra^2G^2}$$

which is the case when the incident signal illuminance is great enough, the function at eqn. (8) approaches the "ideal" function described by eqn. (6). If, however, the signal illuminance is small enough to make

$$\phi_S\sigma p \ll \phi_B\sigma p + i_d + \frac{1}{20Ra^2G^2}$$

then $S/N(\mathrm{dB}) = K + 20 \log(\phi_S\sigma p)$

in unit bandwidth.

Hence from eqns. (3) and (4)

$$S/N(\mathrm{dB}) = K' - 8m_v \text{ in unit bandwidth} \quad \text{......(9)}$$

which is a straight line of slope -8 dB per magnitude.

Equations (6) and (9) describe asymptotes to the operational signal/noise function. The breakpoint in the asymptotic function will be where

$$\phi_S\sigma p = \phi_B\sigma p + i_d + \frac{1}{20Ra^2G^2} \quad \text{......(10)}$$

and at this point the actual function will be 3 dB down on its asymptotic approximation. This is illustrated in Fig. 3.

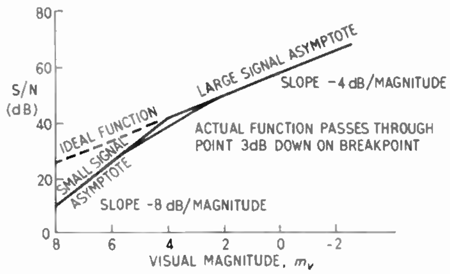


Fig. 3. Operational signal to noise function.

To determine the breakpoint in the asymptotic function it is necessary to examine the quantity

$$\phi_B\sigma p + i_d + \frac{1}{20Ra^2G^2}$$

and for this it is convenient to construct a "breakpoint indicator".

6. Breakpoint Indicator Construction

With a knowledge of the photomultiplier characteristics and its load, and of the amount of light flux from the sky background which falls on the cathode, it is possible to determine the values of S/N(dB) and m_v where the "ideal" signal/noise function of Fig. 2 will break from a -4 dB per magnitude slope into a -8 dB per magnitude slope. If these values are plotted on Fig. 2, the point at which such a plot cuts the "ideal" slope will be the breakpoint.

6.1. *Thermal Noise*

The term $1/20Ra^2G^2$ represents an equivalent cathode current, the shot noise on which is equal to the thermal noise generated in the collector load at a normal temperature. If, as is usually the case, values of G and R can be chosen so that the term is very small compared with i_d, then thermal noise can be disregarded. If the use of a photomultiplier having a small number of stages, and therefore a low value of G, is being considered, then the equivalent cathode current due to thermal noise may be significant compared with i_d. In general, therefore, it is convenient to consider the cathode dark current to have a value i_{de} which is the sum of i_d and the equivalent thermal cathode current. The breakpoint indicator plot is then described by the function

$$\phi_S\sigma p = \phi_B\sigma p + i_{de}$$

6.2. *Dark Current Asymptote of the Breakpoint Indicator*

Suppose the background illuminance is very small so that $i_{de} \gg \phi_B\sigma p$. Then the breakpoint will occur where $\phi_S\sigma p = i_{de}$. In this case

$$S/N(\mathrm{dB}) = 10 \log \left[\frac{F}{2ea^2} \cdot \frac{i_{de}}{2}\right] \text{ in unit bandwidth} \quad \text{......(11)}$$

This indicates a value of S/N(dB) which is independent of m_v, so that under very low background conditions the breakpoint indicator would be a straight line parallel with the abscissa in Fig. 2, and at a level determined by i_{de}. If we put i_{de} equal to 10^{-18} this line is at S/N(dB) equal to zero, and each tenfold increase of i_{de} will raise the level by 10 dB.

6.3. *Background Asymptote of the Breakpoint Indicator*

If the background illuminance is so great that the dark current noise is negligible, then $\phi_B\sigma p \gg i_{de}$. In this case the breakpoint will occur where

$$\phi_S\sigma p = \phi_B\sigma p$$

i.e. $m_v = M$

where M is the equivalent visual background illuminance. ......(12)

(The background illuminance is the product of the sky luminance, I, and the focal plane aperture area, A. Appendix 1 and the sky luminance curves in Fig. 6 show the relationship between I, A, and M.) Equation (12) indicates a value of m_v which is independent of S/N(dB), so that under conditions of very high background illuminance the breakpoint indicator would be a straight line parallel with the ordinate in Fig. 2, and at a value of m_v determined by the sky luminance and the area of the focal plane aperture.

6.4. *Complete Breakpoint Indicator*

Asymptotes to the complete breakpoint indicator have been described which can be superimposed on the "ideal" signal/noise function. A further point on the breakpoint indicator can be determined by considering the case where $\phi_B \sigma p = i_{de}$. In this case, at the breakpoint

$$\phi_S = 2\phi_B$$

whence, applying the relationship between illuminance and visual magnitude (eqns. (3) and (4))

$$-8{\cdot}8703 - 0{\cdot}4\, m_v = 0{\cdot}3010 - 8{\cdot}8703 - 0{\cdot}4\, M$$

or

$$m_v = M - 0{\cdot}7525 \qquad \ldots\ldots(13)$$

Also, at the breakpoint, $\phi_S \sigma p = 2 i_{de}$ which gives a value for S/N(dB) which is 10 log 2 dB above the level of the dark current asymptote of the breakpoint indicator (see eqn. (11)). The breakpoint indicator therefore passes through a point 3 dB above its dark current asymptote, and 0·7525 magnitudes below its background asymptote.

Using similar methods any number of points on the breakpoint indicator can be plotted so that it can be drawn as shown in Fig. 4. If standard scales of S/N(dB) and m_v are used, a cursor defining the shape of the breakpoint indicator is an easily made and useful tool facilitating very rapid assessment of the available signal/noise ratio in a given system.

7. The Relationship Between System Parameters and their Effect on the Available Signal-to-Noise Ratio

Figure 5 represents a complete signal/noise plot of a postulated system, and shows in graphical form the significance of each of the system parameters mentioned in Section 3. The level of the "ideal" function is mainly controlled by p and σ, and to a small extent by F and a. The level of the dark current asymptote of the breakpoint indicator is mainly determined by the photomultiplier cathode dark current, and to a small extent by R, a, and G. The position of the background asymptote of the breakpoint indicator is determined by the focal plane aperture area, A, having due regard to the expected sky luminance, l.

The available signal/noise function cuts the demanded signal/noise level at a value of m_v, which indicates the magnitude of the faintest star which will provide a satisfactory operating signal.

8. Illustration of Use

The usefulness of this method of assessing available signal/noise ratio can best be presented by considering typical design problems.

To achieve the accuracy and information rate required, a given star tracker system demands, from its photomultiplier detector, a signal/noise power ratio of 40 dB. The instantaneous field of view required is such that the focal plane aperture cannot be less than 0·5 square degrees, and the tracker is required to operate outside the atmosphere where the sky luminance is estimated to be about 3×10^{-4} M.K.S. units. The use of two alternative types of photomultiplier is being considered, both having a cathode sensitivity of 50 μA per lumen, the cathode dark currents being 3×10^{-16} amps and 10^{-14} amps respectively. The problem is to determine the minimum size of the optical system entrance pupil which will enable the tracker to operate on stars of $m_v = 3$ or brighter, and how the choice of photomultiplier will affect this size.

The problem is readily solved by constructing a system plot as illustrated in Fig. 5, which shows the case for the low dark current tube. (The case for the high dark current tube has been omitted for clarity.) Reference to Fig. 6 shows that, for a focal plane aperture of 0·5 square degrees and a sky illuminance of 3×10^{-4} M.K.S. units, the equivalent magnitude of background illuminance M is 4·25; this value, and the known values of cathode dark current fix the

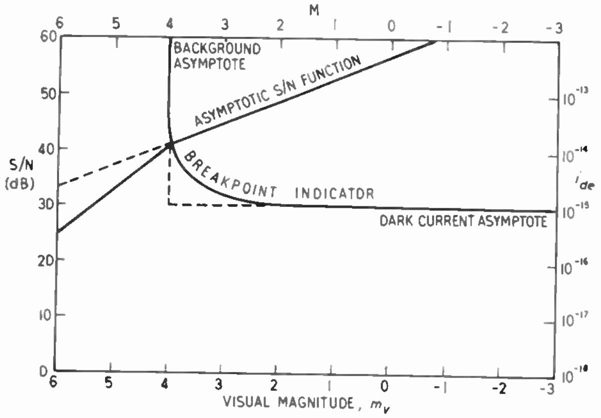


Fig. 4. Breakpoint indicator construction.

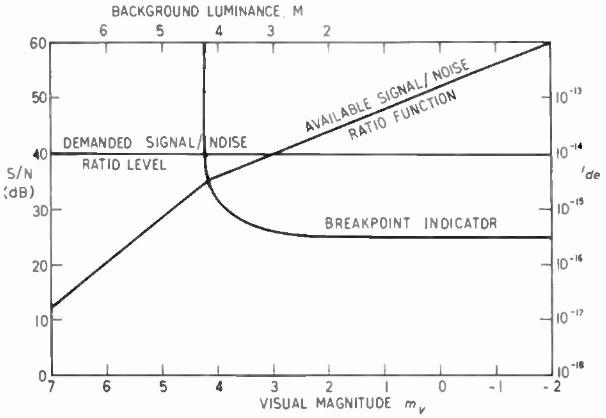


Fig. 5. Available signal/noise ratio plot for a star tracking system.

positions of the breakpoint indicators for the photomultipliers being considered. The available signal/noise ratio function required is one which will cut the demanded signal/noise ratio level at $m_v = 3$. A knowledge of the slopes of the asymptotes to this function enables it to be drawn at the appropriate level. In the case of the low dark current tube, the function cuts $m_v = 0$ at S/N(dB) = 52. Whence, from eqn. (6), and putting $\sigma = 50$ µA/lumen,

$$52 = 94{\cdot}25 + 10\log(50\times 10^{-6}) + 10\log p$$

and $\quad p = 1{\cdot}189$ in^2

If the breakpoint indicator is repositioned to take account of the increased dark current of the alternative tube, it is found that the high signal asymptote of the signal/noise ratio function is raised by about 3 dB; the penalty for using this tube would therefore be to increase the minimum area of the optical entrance pupil by a factor of 10 log 3, i.e. an area of about $2\frac{1}{2}$ in^2 would be necessary.

In some applications, particularly where the specification calls for initial setting up from any arbitrary attitude, design problems arise which are concerned with the way in which stars are distributed. The relationship between star distribution and this type of problem is briefly discussed at Appendix 2.

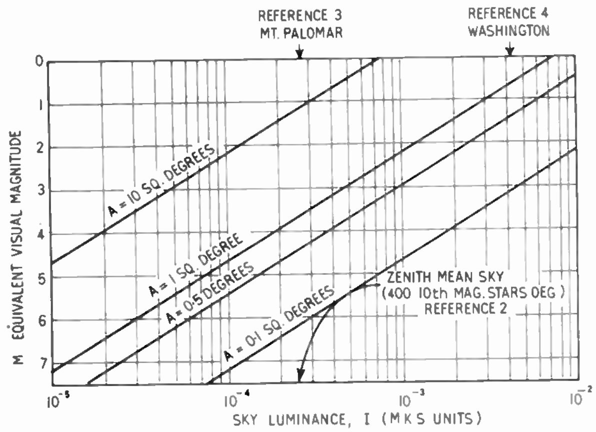


Fig. 6. Night sky luminance.

9. Conclusion

In the initial stages of design or feasibility study, the designer is often more concerned with the relationship between the factors involved than with the precise solution of any particular problem. The study of star tracker systems has so far no established set of parameters and conventions; an attempt has been made in this paper to define at least some of the fundamental parameters common to all systems, and to show the way in which they are related with each other.

10. Acknowledgments

The author wishes to acknowledge with thanks the helpful criticism of the original manuscript offered by Squadron Leader G. H. Pennington, Guidance Squadron, R.A.F. Technical College, Henlow.

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12. Appendix 1

Effective Visual Magnitude of Sky Luminance

A square degrees $= A\left[\dfrac{2\pi}{360}\right]^2$ steradians = the angular area of the focal plane aperture.

I M.K.S. units is the sky luminance.

ϕ_B lumens per square metre is the illuminance through the focal plane aperture.

M is the equivalent visual magnitude.

Then $\phi_B = IA\left[\dfrac{2\pi}{360}\right]^2$ lumens/m^2. Also ϕ_B is related to visual magnitude so that

$$\phi_B = \frac{10^{-6}}{2{\cdot}5^{(M-0{\cdot}8)}}$$

Hence

$$\log I = -6 - 0{\cdot}4\,M + 0{\cdot}32 - \log A + 2\log\left(\frac{360}{2\pi}\right)$$

$$= -0{\cdot}4\,M - \log A + (\bar{3}{\cdot}8764)$$

If A is constant, M as a function of I has a slope of 2·5 magnitudes per decade. M can be plotted as a function of I with A as parameter.

Putting $M = 0$, $\quad A = 1$,

$$\log I = \bar{3}{\cdot}8764$$

and $\quad I = 7{\cdot}523\times 10^{-3}$ M.K.S. units.

If I is constant,

$$\log A = K - 0{\cdot}4\,M$$

i.e. a change of 2·5 magnitudes per decade.

From the above, M is plotted as a function of I with A as parameter in Fig. 6, where the values of night sky luminance derived from references 2, 3 and 4 are indicated.

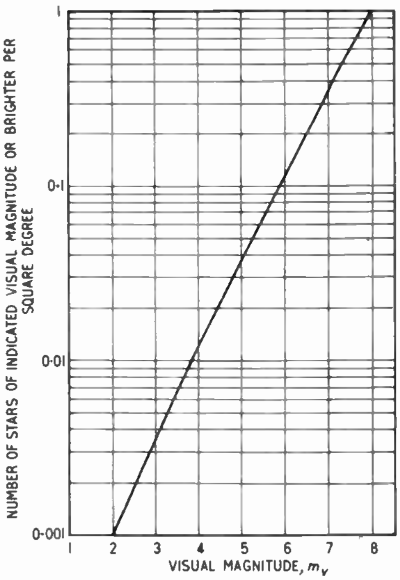

Fig. 7. Mean stellar density (derived from reference 5).

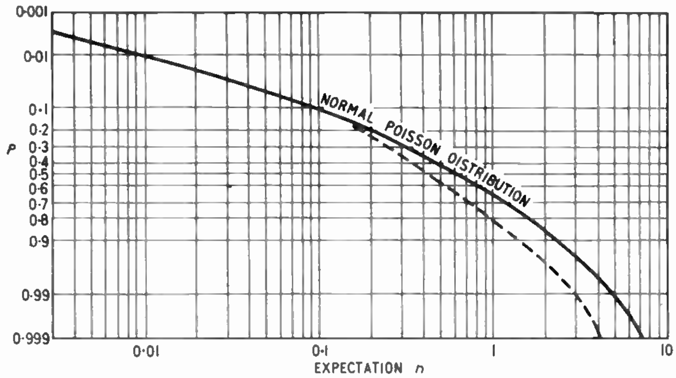

Fig. 8. Probability (P) of at least one star vs. expected number of stars (n).

13. Appendix 2
Star Distribution

The distribution of stars in the celestial sphere is random, and the mean density of stars brighter than a specified visual magnitude is known.[5] In Fig. 7 mean stellar density is plotted as a function of the visual magnitude of the faintest stars in a group.

A star is an isolated event in a continuum of area; in general the distribution of isolated events in a continuum would be expected to follow a Poisson distribution law so that:

$$P = \frac{n^x \, . \, \exp(-n)}{x!}$$

where P is the probability of a number x in an area where the average number is n.

In Fig. 8 the expectation, n is plotted as a function of the probability, P, summed for all values of x greater than 0, the full line being the normal Poisson probability curve. A Poisson distribution is strictly valid only when the expectation and variance are numerically equal, and in reality the Milky Way introduces a disturbing factor which is likely to modify the Poisson law. The broken curve at Fig. 8 is based on a short series of star counts performed by the author and is tentatively offered as a closer approximation to reality than the unmodified Poisson law.

Figures 7 and 8 give an approximate solution to the problem of determining the area of the celestial sphere which must be made available to an optical system in order to find at least one star brighter than a specified magnitude with an acceptable probability of success. Suppose a star brighter than $m_v = 3$ is required. From Fig. 7 the mean density of such stars is 0·0035 per square degree. If 99% probability of success is required, the expectation must be 3 (using the broken curve in Fig. 8). The area which must be made available is therefore

$$\frac{3}{0{\cdot}0035} = 850 \text{ square degrees.}$$

Another problem which is concerned with star distribution is that of ensuring that only one star is included in the instantaneous field of view afforded by the focal plane aperture. Many systems employ a fairly large focal plane aperture area (of the order of one square degree) which introduces a real possibility of interference from an unwanted star. Here the designer is concerned to know how large the focal plane aperture can be without exceeding a given probability that an unwanted star of detectable magnitude will be included in the field of view.

As an example, suppose the detecting system is such that stars fainter than $m_v = 5$ cause an acceptably low level of interference, and that a 1% probability of interference is tolerable. From Fig. 7 the density of stars brighter than $m_v = 5$ is 0·04 per square degree, and from Fig. 8 a probability of 0·01 coincides with an expectation of 0·01 stars. This expectation will be obtained in an area of $\frac{0{\cdot}01}{0{\cdot}04} = 0{\cdot}25$ square degrees. This would be the maximum permissible value for the focal plane aperture area, A.

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A New Gamma Radiation Monitor

By

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Presented at the Symposium on "Electronic Instrumentation for Nuclear Power Stations" in London on 29th March 1961.

Summary: An instrument for continuously monitoring the levels of gamma radiation in the vicinity of a nuclear reactor is described. The monitor employs transistors throughout except for an electrometer valve. Dose rates from 0·1 to 100 mR/hr are measured on a pseudo-logarithmic scale.

1. Introduction

To ensure that people in the vicinity of a nuclear reactor are not subjected to radiation dose rates beyond a safe limit, gamma monitoring equipment is used to indicate and record the radiation level, as well as to provide an alarm when the limit is exceeded. The general background level of gamma radiation in the working area of a nuclear power station is expected to be between 0·1 and 1 milliroentgen per hour. The monitor described is designed to cover 0·1 to 100 mR/hr in one range, thereby allowing for a hundred-fold increase in the radiation level. By eliminating the necessity for range switching the operation of the instrument is considerably simplified and by designing the overall response to be such that the lowest decade occupies about half the scale length, while the higher decades cover the remaining scale logarithmically, the dose rates of immediate interest are easily measured. The mechanical design is such that the instrument is watertight, transportable and self-contained, requiring only external mains supply.

A direct current proportional to the intensity of the incident radiation is produced by the ionization chamber. This current is amplified by a circuit consisting of an electrometer triode, used as a logarithmic element, followed by a transistor chopper-type d.c. amplifier, as shown in Fig. 1. A high dose rate alarm operates a warning lamp on the instrument and the trip level can be adjusted to any point in the instrument range. Facilities are provided for operating an external recorder and alarm. Built-in test facilities are provided for routine testing of the instrument.

A stabilized power unit provides the low voltage supplies. The chamber polarization voltage is stabilized by a circuit using the avalanche breakdown characteristic of a normal silicon rectifier to produce a high voltage stabilizer.

2. The Ionization Chamber

When ionizing radiation enters an ionization chamber, the ions produced are swept continuously to the positive or negative electrodes by the polarizing electric field and a steady current flows proportional to the intensity of the incident radiation. To minimize the measurement problem the current output from the chamber should be as large as practicable.

As this current is dependent on the volume and pressure of the chamber gas filling, a large high pressure chamber is used in this monitor having a volume of 4·7 litres and a filling of argon and nitrogen at a total pressure of 20 atmospheres. With this filling a relatively high current sensitivity of 10^{-8} amps/roentgen/hour is achieved. The energy response of the chamber is such that substantially true dose rate measurements are maintained with radiation energies from 0·2 to about 3 MeV. Below this energy wall effects in the chamber produce distortion in the chamber energy response. The response to high energies has not yet been determined, but it seems likely that the chamber could be used for gamma energies up to 5–10 MeV.

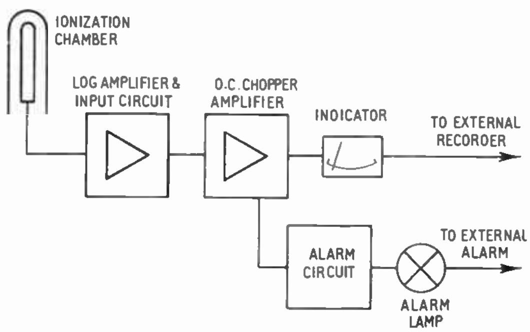

Fig. 1. Block diagram of the gamma radiation monitor.

The chamber polarization voltage is chosen so that almost complete collection efficiency is maintained over the range of the instrument.

3. The Input Stage and Test Circuit

The current produced by the ionization chamber is small (10^{-12}A at 0·1 mR/hr), and as at present no solid-state device with the necessary sensitivity is available, a thermionic valve is essential at the input stage, shown in Fig. 2.

The logarithmic element (V1) is a directly heated electrometer triode with the chamber current feeding directly into the grid. The grid and cathode can be considered as a diode operating in the retarded field

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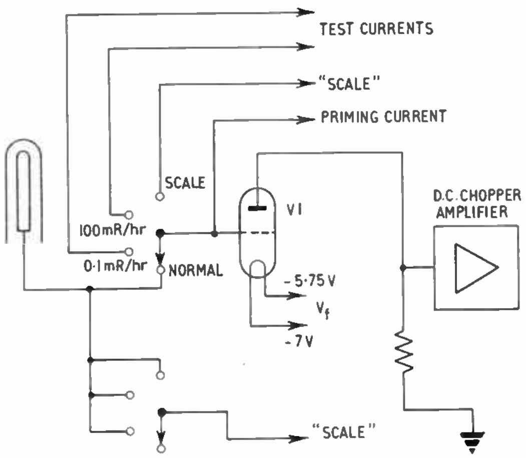

Fig. 2. Logarithmic amplifier and test circuit.

condition, so that a logarithmic relationship exists between the grid current and voltage. The grid voltage controls the anode current of the valve in the normal way. Using an electrometer triode in this stage has several advantages over the normal pentode or thermionic diode circuits. The larger mutual conductance of the triode compared with that of the pentode provides higher signal levels for the succeeding transistor amplifier and so minimizes drift in these stages.

The triode has the important advantage over a thermionic diode, used as a logarithmic element, that whereas in a diode the logarithmic relationship between the diode current and voltage is inherently sensitive to variations in the cathode temperature, in the triode an inherent cancelling effect exists, tending to maintain a constant anode current with changes in filament temperature.[1] A decrease in the filament voltage tends to reduce the anode current, since the cathode emission is lowered. The ionization chamber current in the grid does not change, so the grid potential increases to satisfy the grid-cathode diode relationship and this increase tends to restore the anode current to its initial value. If this cancelling effect were perfect, absolute stabilization would be possible, but as this is usually not the case, heater voltage stabilization is required.

Any logarithmic element has a range beyond which its characteristic is no longer truly logarithmic. For the triode used in this circuit (ME 1404) a range of three decades is quoted (3×10^{-12} to 3×10^{-9}A). It has been found that currents down to 1×10^{-12}A can be measured with this triode without substantial divergence from the logarithmic characteristic. If some distortion of the response can be tolerated currents greater than 1×10^{-9}A can be measured.

To prevent the grid voltage of the valve from building up in the absence of radiation and to swamp the effects of spurious grid currents, a small priming current, equivalent to a dose rate of 0·05 mR/hr or 5×10^{-13}A, is applied to the grid of the electrometer valve by a resistor to which a preset constant voltage is applied. Any increase in the grid voltage of the valve would produce a rise in the grid to cathode impedance, which in turn would result in increased time constants with the inherent chamber and stray capacitances; considerable delay in indication could thus arise should a sudden change in radiation level occur.

The input of the electrometer valve, which is normally connected to the ionization chamber, can be switched to test positions, see Figs. 2 and 3. In two of the test positions built-in current generators simulate the ionization chamber currents corresponding to dose rates of 0·1 mR/hr and 100 mR/hr and should produce equivalent meter indications. If any drift has occurred in the amplifier, the instrument can be readjusted with the ZERO and calibration (CAL) controls (see Fig. 4), where 0·1 mR/hr is set with ZERO and 100 mR/hr with CAL control. For routine testing therefore an external gamma source is not required. The ion chamber is disconnected during these tests and checking can be carried out in the presence of external radiation.

The main cause of drift will be due to small inherent variations of the zero signal grid potential of the electrometer valve. By operating the monitor con-

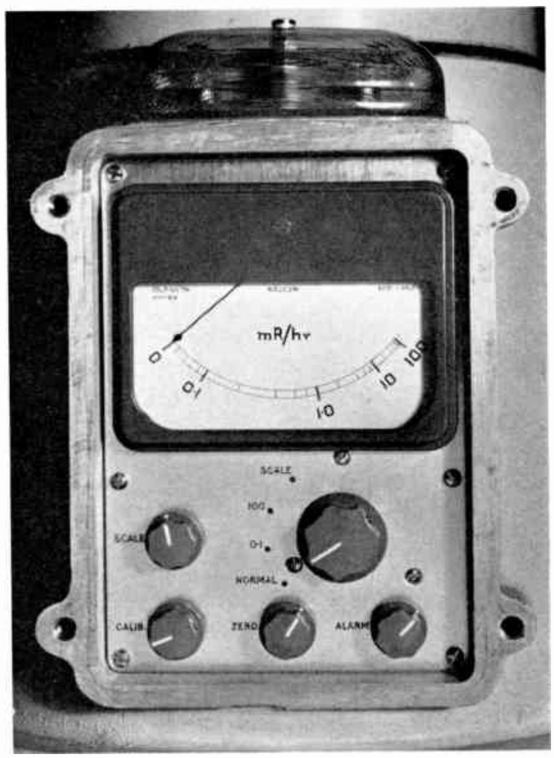

Fig. 3. Control panel.

tinuously, the electrometer grid potential can reach a state of equilibrium and so minimize such a drift.[1]

Leakage of the ionization chamber filling, which would result in a lower sensitivity, is expected to be less than 1% per year.

The current generators included in the monitor use high value resistors, which may drift slightly with time, and hence some long-term variation in the simulated ion chamber currents and priming current may occur. Only the latter is operative during normal measurements. A 10% change in the priming current will produce a negligible error of about 1% at the normally expected dose rate of 0·5 mR/hr.

To detect any long-term drift in the simulated chamber currents periodic checks, say every 6 months, with a known gamma source are suggested. Any error in these currents can be corrected by readjusting potentiometers in the monitor (see Section 4).

Due to the log characteristic of the valve any ripple voltages superimposed on the grid will result in a mean level change in the output. Thus ripple on the ion chamber supply line is transferred by the chamber inter-electrode capacitance to the grid of the valve and can produce an error. The design of the power supply, however, results in a ripple on the ion chamber h.t. of about 5 mV peak-to-peak which represents a negligible change in the mean level of around 1% per decade of dose rate.

To set the alarm trip level in the monitor the grid of the electrometer valve is connected, in the alarm test position SCALE, to an adjustable voltage source. Any required grid voltage can be selected and hence any radiation intensity in the range of the instrument simulated and displayed on the meter. With this facility and the ALARM set control the alarm trip level can be set and checked.

In all the test positions of the input switch the output from the chamber is connected to the adjustable voltage source. This ensures that during the tests the ionization chamber does not charge up to a potential that may damage the valve when the switch is returned to the normal position. There is an advantage in connecting the chamber output to this voltage during these tests, instead of to the cathode of the valve, in that when switching the chamber back to the grid it will have a potential that is close to the normal grid operating voltage. The result is a reduction of the recovery time of the grid-cathode system and hence in the recovery time of the indicated reading.

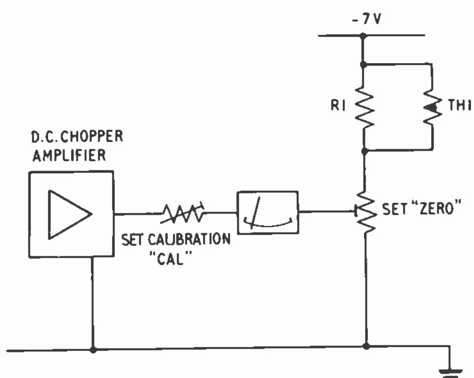

Fig. 4. Output circuit.

The insulation of the input stage is important. To minimize any leakage currents a dry airtight box houses this circuit.

4. Chopper-type D.C. Amplifier

The anode current of the electrometer valve must be amplified before it can be displayed on the indicating meter. To achieve low drift in the instrument, a transistorized d.c. chopper amplifier is used. The input chopper transistor converts the d.c. input to a proportional a.c. signal, which is then amplified by an inherently drift-free a.c. amplifier. The a.c. output from the amplifier, after rectification, drives the indicating meter, and is also used directly to operate the high level alarm. The valve anode current necessarily contains a large standing current which must be backed off, so that the meter displays only its variable part; this is achieved at the amplifier output by the ZERO control.

The chopper transistor does not behave as a perfect switch, because when bottomed a residual collector-emitter voltage exists and when cut off a leakage current flows. These effects are temperature sensitive, but are greatly reduced by using the transistor in an inverted mode.[2] By applying an optimum base drive of about 1 mA to the chopper transistor, the variation with temperature in the residual voltage drop across the transistor is further reduced. Hence the error signal when the chopper transistor is bottomed will appear as a constant input to the amplifier and can be backed off at the output. To reduce the error signal when the chopper is cut off, a low leakage current transistor must be used. As silicon transistors have lower leakage currents than germanium types, a silicon transistor is most suitable for this simple type of chopper. The disadvantage of a larger collector-emitter voltage in a silicon transistor is not important, as it only constitutes a constant input which can be backed off at the amplifier output.

The drive to the chopper transistor is produced directly from the secondary of the power supply mains transformer. (See Fig. 5.) A clipping diode D1 is used to limit the positive half cycle of the 50 c/s sine wave to +1 V, so avoiding excessive positive drive, which could result in damage to the transistor. Variations in the mains voltage (+8%, −20%) produce no appreciable error in the meter indication, as the optimum base drive is not very critical.

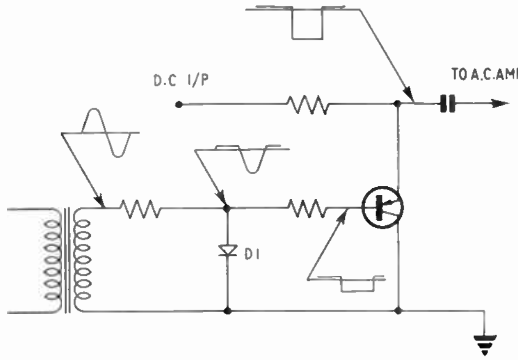

Fig. 5. Transistor chopper.

The modulated input signal is amplified by a three stage a.c. amplifier, consisting of emitter follower input and output stages, with a single stage of voltage amplification. Overall feedback defines the amplifier gain and makes it virtually independent of transistor parameters. A voltage doubler rectifying circuit is used as the demodulator.

The indicating meter (1 mA f.s.d.) has a semi-logarithmic movement, which together with the log response of the input stage, produces the output characteristic of the monitor. Thus the first decade 0·1 to 1·0 mR/hr occupies approximately half the scale. This is usually of greatest interest as the normal dose rate in the reactor building and their surroundings is expected to be in the region of 0·5 mR/hr. The other decades are displayed logarithmically on the remaining scale length. A resistor, in series with the meter movement, provides an output suitable for driving an external 100 mV recorder. By deriving the semi-logarithmic characteristic from a meter movement rather than the more usual diode feedback circuits, the overall response of the instrument is made less dependent on temperature and supply changes, as well as diode characteristics.

The meter scale, although basically logarithmic, has a zero position which corresponds to an input of the standing priming current (5×10^{-13}A). This current is made up of an adjustable priming current and any unknown leakage currents in the input stage. This makes the zero adjustment independent of a knowledge of the leakage currents. The design of the input housing ensures that any possible leakage currents are reduced to a minimum.

The true ionization chamber current is superimposed on the priming current so that some distortion of the basic log response occurs, as a dose rate of 0·05 mR/hr must be added to each decade. Clearly this factor becomes progressively smaller with the higher decades, i.e. it represents 5% of the first decade 0·1 to 1 mR/hr, 0·5% for the second decade, etc. This distortion does not, however, produce any inaccuracy in the dose rate indication as it has been taken into account in the design of the meter scale.

To calibrate the monitor a known radioactive source is used to produce dose rates corresponding to the cardinal points on the meter and the zero point and the sensitivity of the meter circuit are adjusted, as shown in Fig. 4. With this method any production spread in the gain and the standing current of the electrometer triode is compensated. The calibration current generators, which consist of high value resistors to which adjustable stabilized voltages are applied, are then set up to provide the test currents equivalent to the dose rates 0·1 and 100 mR/hr. They are adjusted to correspond to the dose rate indications on the output meter produced by the known source, thereby making the test position accuracy dependent only on that of the source.

The temperature stability of the electrometer valve supply and the design of the input chopper circuit are such that they produce a negligible error with change in ambient temperature. An a.c. device is used for the main transistor amplifier so that small variations in the d.c. biasing, produced by temperature drift, are not important. The demodulator diode temperature coefficient is -3 mV/deg C approximately, and this is the main source of temperature drift, when the instrument is operating normally. This drift causes an increase in the meter indication, so that, by producing an equal increase in the backing-off voltage at the output meter, compensation is possible. The simplest method of compensation is produced by inserting a thermistor TH1 (see Fig. 4), into the backing-off network. The negative temperature coefficient of TH1 is linearized and adjusted to the required value by shunting it with a resistor of negligible temperature coefficient (R1). Knowing the required resistance variation of the shunt and series resistance combination to produce compensation, the values of the components can be found.[3]

5. High Dose-Rate Alarm

The alarm is an a.c. circuit driven directly from the a.c. amplifier (Fig. 6). This method has the advantages of low backlash and of being independent of temperature errors in the demodulator. The trip level of the alarm, which is continuously variable over the range of the instrument, will operate the alarm lamps when the incident dose rate exceeds the preset level. Facilities are provided on the alarm relay for a remote alarm indication. The alarm is basically a bistable circuit (VT1, VT2) but operates as a monostable circuit with an a.c. drive. The d.c. bias is arranged so that in the non-alarm condition VT1 is cut off. The a.c. drive for the alarm is taken from potentiometer RV1, which sets the trip level, and as the a.c. amplifier output increases, the drive to VT1 increases until the

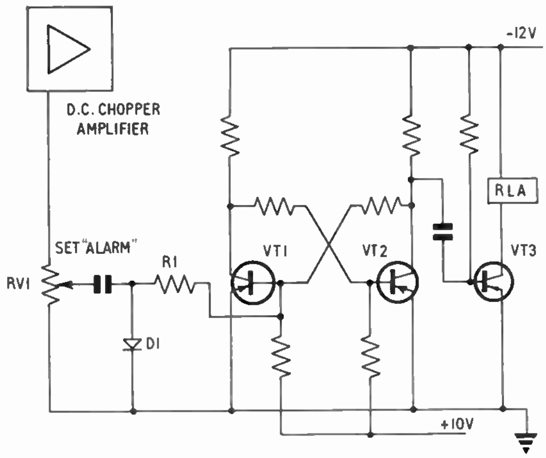

Fig. 6. High dose-rate alarm.

circuit is triggered into oscillation. The relay switching transistor VT3 is conducting in the non-alarm condition. When the alarm circuit is triggered into oscillation the positive-going pulses from the collector of VT2 cut off VT3, relay RLA becomes de-energized, and alarm lamps are operated. Two lamps connected in parallel are used to indicate an alarm. Should one lamp fail alarm indication is still possible. If VT3 goes open circuit, a permanent alarm condition arises. Transistor failure in the monostable circuit could prevent an alarm indication. This tendency has been minimized by under-rating these components.

6. The Power Unit

This provides the stabilized voltages for the circuits, the chopper transistor base drive, and the power for the alarm lamps. A series stabilized supply, shown in Fig. 7, provides the low voltages (A). Stabilization is produced by the control transistor VT1 in series with the supply and the load. A fraction of the voltage across the load is compared with a reference voltage produced by the Zener diode D1 and any difference in these is used to control the voltage drop across the series element. This control produces a constant voltage across the load. To minimize changes in the reference voltage due to input voltage or load variations, the Zener diode D1 is operated at a low dynamic impedance point of its characteristic. The required bias is produced by a large standing current from the stabilized line.

As the ambient temperature increases, the transistor parameters V_{be} and I_{co}, and the reference voltage of D1 will vary. All these variations, except those due to the base emitter voltage of VT3 and the reference diode, are corrected by the feedback loop. The drift in V_{be} of transistor VT3 can be balanced by choosing a Zener diode with an equal and opposite temperature coefficient. The Zener diode OAZ 203 has a temperature coefficient of approximately $+2{\cdot}5$ mV/deg C at 10 mA. This balances the V_{be} drift of the OC72 which is used for VT3 ($-2{\cdot}5$ mV/deg C).

The stabilized electrometer valve heater supply (V_f) is taken from a potential divider across the series stabilized line (A) and is controlled by a second stage of stabilization produced by the diode D2 and D3. The variation with temperature in the forward characteristics of a silicon rectifier D3 is used to cancel that in the characteristic of the Zener diode D2 to produce optimum temperature stability at the required reference voltage. A resultant drift of +5 mV (20 to 50° C at 6·7 V, 20 mA Zener current) is obtained.

A stabilized voltage of opposite polarity is produced by a simple d.c. converter VT4 (B).

The ionization chamber polarization voltage supply is stabilized by a very simple yet novel method using the avalanche breakdown characteristic of a normal semiconductor rectifier. This breakdown mechanism is identical to that in Zener diodes at stabilization voltages greater than about 6 volts. The reverse characteristic of a semiconductor rectifier shows a rapid decrease in dynamic slope resistance when avalanche breakdown occurs, and this effect can be used to produce a stabilizing element. The breakdown voltage of a silicon diode is generally larger than that of a germanium type, and the change in the dynamic slope resistance at the breakdown voltage point appears to be greatest in low-power diodes. Hence low-power silicon rectifiers appear to produce the best avalanche stabilizers, but as the avalanche voltage is of the order of several hundred volts, the current that can be controlled is necessarily small.

The avalanche diode can be used in the standard types of d.c. stabilization circuits to provide a reference voltage; in this instrument, however, a.c. stabilization is used as it produces optimum stabilization with the minimum number of components. In the basic a.c. regulator circuit (Fig. 8) when an alternating voltage is applied to the avalanche diode it limits the positive and negative half cycles. The diode conducts normally when the voltage is in the negative half-cycle, but in the positive half-cycle the diode will

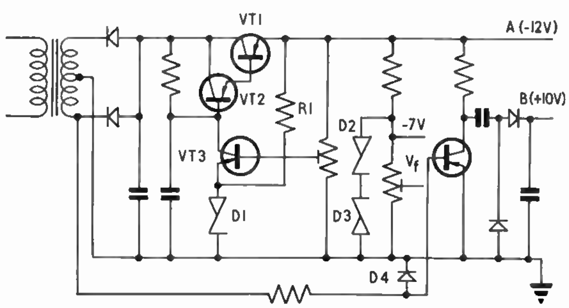

Fig. 7. H.T. stabilized power supply.

not conduct until the applied drive exceeds the avalanche voltage. This results in a stabilized drive which can produce a stabilized d.c. output. The series resistance R5 limits the current through the diode.

If a series capacitor C1 is included in the basic a.c. regulator, as shown in Fig. 9, the drive available to D1 is increased. For a given stabilization therefore, this circuit requires a lower voltage $\hat{E}$ than the basic a.c. regulator of Fig. 8.

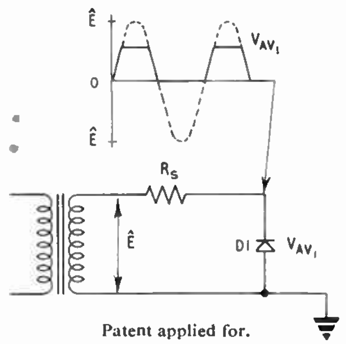

Fig. 8. A.C. regulator with avalanche diode.

The diode D1 is not a perfect stabilizer and mains voltage variations could produce some change in the d.c. output, so that any transients would transfer their charge, via the chamber inter-electrode capacitance, to the grid of the electrometer valve. This effect is considerably reduced if a diode with an avalanche voltage slightly lower than that of D1 is used for D2, because then a rapid discharge path is available if C2 is charged to a higher voltage than normal. During the positive half cycle D2 is forward biased and C2 will charge up to a voltage defined by D1. In the negative half cycle D2 is reverse biased with a voltage exceeding its avalanche voltage and so C2 will tend to discharge to a voltage defined by D2. Equilibrium is reached when the charge put into C2 in the positive half cycle is equal to that removed in the negative half cycle and as the time in each cycle during which the output is defined by D2 is longer than that defined by D1 the output voltage is approximately equal to the avalanche voltage of D2. For optimum stabilization the breakdown voltage of D2 should be within a few volts of that of D1. This simple method produces a stabilization which compares favourably with more complex circuits. The stabilization of the circuit (Fig. 9) for +10%, −20% mains variations is +0·05%, −0·1%. The temperature coefficients of these diodes, in the avalanche condition, appears to contribute approximately +0·1%/deg C variation.

An elaboration of the normal RC filter is used with the polarization voltage supply to reduce meter fluctuations due to mains supply transients. This requires a long time constant in the h.t. output of the

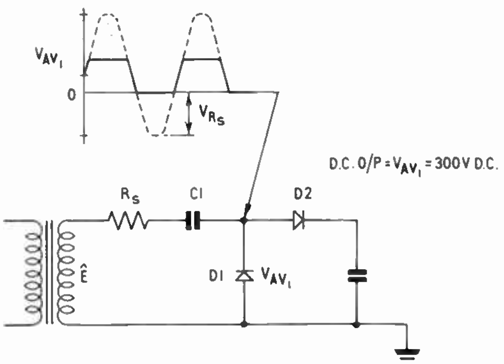

Fig. 9. Stabilized ion chamber supply.

power supply; the basic circuit is shown in Fig. 10 (*a*). On switching the monitor on, however, this circuit would also result in a long delay, of the order of minutes, before C1 could be fully charged and an accurate reading of dose rate could be taken. This is due to the charging process producing an increasing h.t., which is transferred to the grid of the electrometer valve via the chamber interelectrode capacitance, causing an upward drift in the meter indication. This delay is substantially reduced by using a voltage-conscious resistor in the series element. The Zener diode D1 is used as such a resistor in Fig. 10 (*b*). It acts as a two-state device in that when the voltage drop across it exceeds its Zener voltage it represents a low resistance of a few ohms, but as this voltage drop decreases to less than its Zener voltage it becomes a very high resistance in the order of 100 megohms. If C1 is considered to be initially discharged, then when the instrument is switched on, the output voltage from the stabilized supply (V_s = 300 V approximately) will be across R2 and D1. If the Zener voltage of D1 is about 6 V then C1 will charge up with a time constant R_2C_1. The value of R2 is chosen as 10 MΩ and serves to limit the Zener current in D1; if C1 is 1 μF, then the time constant will be 10 seconds. When the voltage drop across D1 decreases below about 6 V, D1 will act as a high resistance and the time constant now becomes much larger thereby reducing the fluctuations due to mains transients. This filter also reduces the ripple from the supply to 5 mV

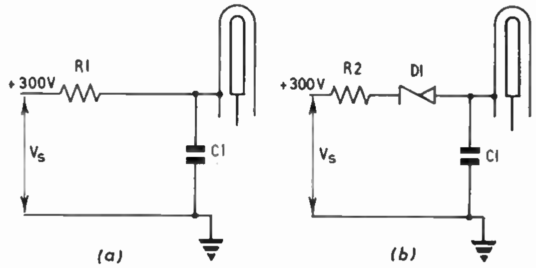

Fig. 10. RC filter for ion chamber supply.

p.p. and so minimizes any standing error in indication due to the log characteristic of the electrometer valve.

7. Conclusion

The Gamma Radiation Monitor is designed primarily for nuclear power stations, where the basic need is for a robust reliable dose rate meter which requires the minimum of adjustment and with which measurements can be quickly and easily taken. The rugged waterproof housing of the monitor (Fig. 11) provides an instrument for industrial as well as laboratory environments. The use of semiconductors wherever possible should result in a long operating life and maximum reliability. Available data on electrometer valves indicate that under continuous operation a life expectancy of 10 years is not unreasonable. The pseudo-logarithmic scale of the monitor is designed to provide clear indication of the dose rates expected in the reactor working area, while higher radiation levels are still indicated without range switching. The test facilities provided in the monitor permit routine checks to be carried out *in situ*, without the necessity of external test gear. The high dose-rate alarm can be set to operate at any point in the range of the monitor, and remote indication of alarm and dose rate are provided. Component or power failure in the measuring instrument would generally be indicated as a zero or less than zero output on the meter or remote potentiometric recorder. As the radiation levels around the reactor are expected to produce considerable deflection on the scale of this monitor the failure conditions can readily be detected.

Table 1 shows prototype performance figures.

Table 1

Overall accuracy	Within ±10% for gamma radiation energizes down to 0·2 MeV.
Stability	Indication change ±3%, ±0·02 mR/hr† for mains change of ±10%. Reading remained within ±1% f.s.d. for one month. Indication changed by ±0·1%/deg C up to temperature of 45° C.
Alarm	Backlash from 0·5% to 1% f.s.d.

† Stability can be ± 3% of indication. To this a factor of ± 0·02 mR/hr is added. The factor is only apparent at low dose-rates.

While the circuitry of the monitor is generally standard the ionization chamber polarization supply represents a simple yet novel method of providing a stabilized high voltage. Although the power available from this type of circuit is small, nevertheless it has many applications particularly in nuclear instruments.

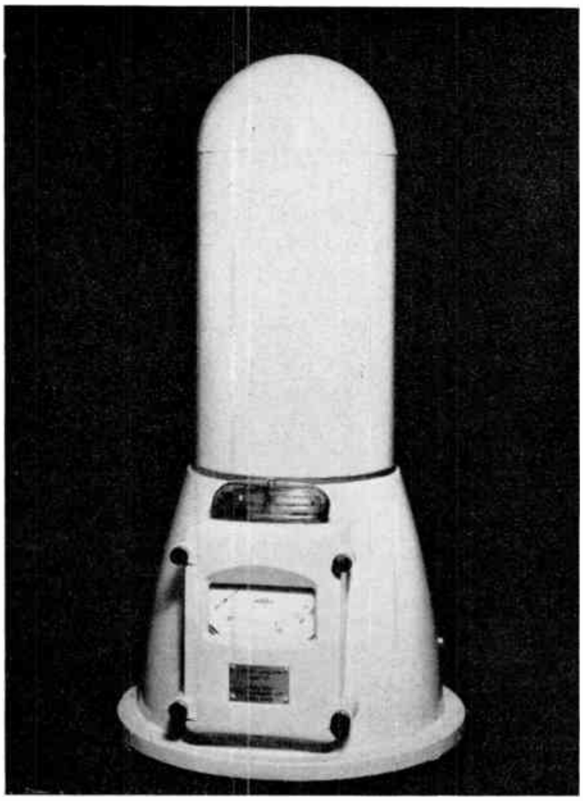

Fig. 11. Gamma radiation monitor.

Development by the semiconductor manufacturers to provide more information on the avalanche characteristics of silicon rectifiers would be useful, as these provide more economical stabilizers than any high voltage Zener diodes available at present.

8. Acknowledgments

The author would like to thank the Management of Marconi Instruments Limited for their permission to publish this paper, and also all those working at Marconi Instruments who have assisted in the design and construction of the instrument.

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News from the Sections . . .

South Western Section

An account of the engineering aspects of the use of transistors for v.h.f. receivers and transmitters was given at the meeting on 20th December by Mr. D. Carey, B.Sc. After dealing with the maximum achievable power outputs from alloy junction types, Mr. Carey went on to show how better results from alloy diffused germanium and silicon types can give outputs of 100 W at a few megacycles or a few watts at some hundred megacycles or so. Epitaxial diffused types seem likely to yield even better results.

The main space consumers for small transmitter layouts are the coils, but the latest innovation of ferrite bead chokes may obviate even this coil disadvantage. The overall efficiency of about 35% for 1 W output at 100 Mc/s compares favourably with larger permanent installations.

In receiver performance, particularly for such obvious uses as Band III operation, transistor receivers are now very satisfactory. Typical overall efficiencies are about 30%–40% for 1·5 W receiver requirement. The most critical stage from the point of view of noise is, as expected, the mixer. At the present stage of the technique, separate local oscillators are being employed. Noise figures of 10 to 15 dB in a 5 kc/s bandwidth were quoted, with even lower figures possible using epitaxial types.

D. R. M.

North Western Section

At the meeting on 7th December Mr. K. F. Slater of the Royal Radar Establishment gave a lecture on "Radar for Civil Aviation Purposes". Firstly he explained the basic information requirements of an air traffic control system, pointing out the need for a surveillance or "feedback" channel. At present, in the United Kingdom, this requirement is met in the airways by the pilot reporting at radio beacon markers, but if the density and flexibility of the system is to be increased then radar must fulfil this role.

Radar requirements and the influence of choice of wavelength on such aspects of "solid" elevation cover, angular discrimination and weather clutter were then discussed. It was seen that, apart from the latter aspect, the use of short wavelengths (10 cm) was advantageous and with currently available techniques of circular polarization and pulse compression the weather clutter could be reduced. A "compressed" pulse is generated by passing a pulse with linear frequency modulation over the pulse duration time through a network with a group delay inversely proportional to frequency. This technique enables high performance radar to operate with effectively short pulses without impracticably high peak powers.

Mr. Slater then discussed height finding and the advantages of a stacked beam system for elevation cover mentioned. Primary radar is the main contribution to the information flow for traffic control but increasing attention is being paid to secondary radar. However, with secondary radar a ground interrogator sends out a suitably coded pulse sequence which is received by the aircraft and used to trigger an airborne transmitter, preferably on a different frequency, which emits a coded reply. Secondary radar gives freedom from ground and other clutter, and a signal independent of aircraft size. At present, trials are in progress in which aircraft can reply automatically to particular interrogation codes with replies giving height and identity.

F. J. G. P.

Scottish Section

"Nuclear Reactor Safety Circuits" was the title of a paper read by T. M. Dowell, B.Sc., A.R.C.S.T., of Hunterston Nuclear Generating Station, in Edinburgh and Glasgow on 10th and 11th January.

In his introduction, Mr. Dowell explained that the term "safety" in this context must imply a factor of an extremely high order.

He reviewed briefly the Hunterston plant as a whole and the general principles on which it operates. The two reactors are of the Calder Hall type, using natural uranium fuel with a graphite moderator and carbon dioxide cooling.

The core of each reactor has 3288 vertical channels each containing 10 fuel elements. The rate of fission and therefore of heat production is controlled by neutron-absorbing boron steel rods suspended in the core. To shut down the reactor, the rods are allowed to fall by gravity into the core.

The various safety devices operate a guard line of relays. Since one guard line in itself might be subject to spurious operation due to component failure, three lines are provided, and it is necessary for relay contacts on two lines to open before the control rods are released and the reactor shut down.

Safe operation of a reactor is limited by the maximum allowable temperature within the core. Owing to the large thermal capacity of the core however, safety measures must take into account trends as well as actual temperatures. This is most readily achieved by monitoring the neutron flux producing fission. The mean current ionization chamber and the pulse counter are used to cover the wide range necessary.

Resistance thermometers are not suitable for temperature measurement because of the effects of radiation on resistivity. Thermocouples have proved to be satisfactory and 1000 are used at Hunterston, each continuously monitored. An elaborate system of grouping the thermocouples and measuring their outputs is employed to detect local rises of temperature, which might produce instability of the reactor.

W. R. E.

U.D.C. 621.372.512.3

Terminal Impedance and Gain of the Series-to-Parallel Transitionally-Coupled Circuit

By

G. J. A. CASSIDY, B.E.E.†

Summary: An examination is made of the terminal impedances and gain of the network consisting of a series-tuned circuit transitionally coupled to a parallel-tuned circuit, to show the usefulness and limitations of the singly-loaded form of the circuit for coupling a transmission line to a vacuum-tube grid or anode.

1. Introduction

During a study of the coupled circuit of Fig. 1, having a series-tuned circuit inductively coupled to a parallel-tuned circuit, in which some of the circuit relationships of Stracca[1] were obtained, the terminal impedances were derived for the case with transitional coupling (which gives maximally-flat, or Butterworth response).[2, 3]

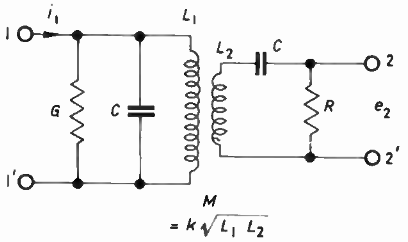

Fig. 1. The circuit configuration under examination. G and R include the terminating impedances.

This form of circuit is useful when it is desired to couple, over a particular frequency range, a resistance of relatively low value to a high resistance which may be shunted by capacitance. This is often the case when a transmission line is to be coupled to the grid or anode of a vacuum tube.

2. Circuit Relationships for Transitional Coupling

The circuit relations equivalent to those of Stracca, for maximally-flat response, for Fig. 1, are

$$L_1 C_1 \omega_0^2 = 1 \qquad \text{......(1)}$$

$$L_2 C_2 \omega_0^2 (1-k^2) = 1 \qquad \text{......(2)}$$

$$1/(1-k^2) = 1 + \tfrac{1}{2}(a_1^2 + a_2^2) \qquad \text{......(3)}$$

where ω_0 is the angular frequency at maximum response,

k is the coefficient of coupling $\{= M/(L_1 L_2)^{\frac{1}{2}}\}$,

$a_1 = \omega_0 L_1 G = G/\omega_0 C_1$ is the shunt-side dissipation factor,

$a_2 = \omega_0 C_2 R = R/\omega_0 L_2(1-k^2)$ the series-side dissipation factor, and

$$a = a_1 + a_2$$

These relations give a transfer impedance (of similar form to the transfer function of Stracca) from terminals 1,1′ to 2,2′ (Fig. 1).

$$|Z_t|^2 = 4\frac{a_1 a_2}{a^4} \cdot \frac{k^2}{1-k^2} \cdot \frac{R}{G} \bigg/ (1+\varepsilon^4) \qquad \text{......(4)}$$

$$= 4\frac{a_2}{a^4} \cdot \frac{k^2}{1-k^2} \cdot \frac{R}{\omega_0 C_1} \bigg/ (1+\varepsilon^4) \qquad \text{......(5)}$$

where
$$\varepsilon = \frac{\sqrt{2}}{a} \cdot \left(\frac{\omega_0}{\omega} - \frac{\omega}{\omega_0}\right)$$

$$= \frac{\omega_0}{\Delta\omega} \cdot \left(\frac{\omega_0}{\omega} - \frac{\omega}{\omega_0}\right)$$

and the 3 dB bandwidth, $\Delta\omega$, is $a\omega_0/\sqrt{2}$

The response, shown diagrammatically in Fig. 2, is symmetrical about ω_0 on a logarithmic frequency scale.[4]

It should be noted that both terminating impedances are included within the circuit for which this transfer impedance has been obtained.

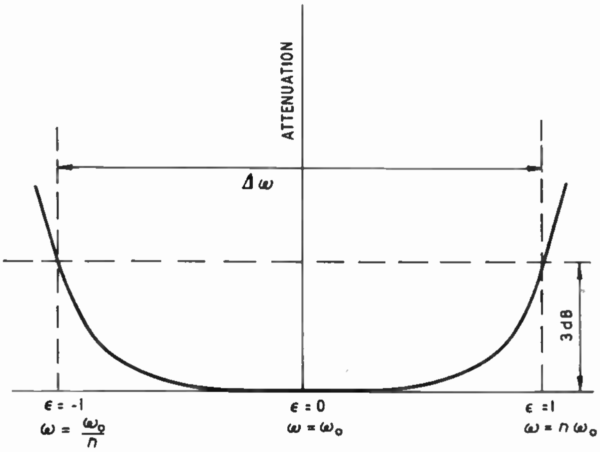

Fig. 2. The form of a "Butterworth" or "maximally-flat" type of response curve.

† Commonwealth Scientific and Industrial Research Organization, National Standards Laboratory, Division of Electrotechnology, Sydney, New South Wales.

3. Terminal Impedances

The terminal impedances may be derived by ordinary circuit analysis techniques. In each case the transformer is taken with its element values determined by the above relations for maximally flat response, the input-side resistive loading is open-circuited and the impedance presented at that terminal pair by the transformer with its design output-side loading is investigated to determine the load presented to the source. The results may be expressed in terms of the design source-resistance and dissipation factors.

The series-side terminal impedance is then

$$Z_2 = \frac{R}{a_2}\,\frac{a_1\dfrac{a_1^2+a_2^2}{2} - \mathrm{j}\left(\dfrac{\omega_0}{\omega}-\dfrac{\omega}{\omega_0}\right)\left\{\left(\dfrac{\omega_0}{\omega}-\dfrac{\omega}{\omega_0}\right)^2+\dfrac{a_1^2-a_2^2}{2}\right\}}{\left(\dfrac{\omega_0}{\omega}-\dfrac{\omega}{\omega_0}\right)^2+a_1^2} \qquad \ldots\ldots(6)$$

At maximum response ($\omega = \omega_0$), this reduces to

$$Z_{2m} = R\,\frac{a_1^2+a_2^2}{2a_1 a_2}$$

The shunt-side terminal admittance is

$$Y_1 = \frac{G}{a_1}\,\frac{a_2\dfrac{a_1^2+a_2^2}{2} - \mathrm{j}\left(\dfrac{\omega_0}{\omega}-\dfrac{\omega}{\omega_0}\right)\left\{\left(\dfrac{\omega_0}{\omega}-\dfrac{\omega}{\omega_0}\right)^2+\dfrac{a_2^2-a_1^2}{2}\right\}}{\left(\dfrac{\omega_0}{\omega}-\dfrac{\omega}{\omega_0}\right)^2+a_2^2} \qquad \ldots\ldots(7)$$

At maximum response ($\omega = \omega_0$) this reduces to

$$Y_{1m} = G\,\frac{a_1^2+a_2^2}{2a_1 a_2}$$

The inverse relationship between these expressions is immediately apparent. It will be further seen that the "matched" condition ($Z_{2m} = R$, or $Y_{1m} = G$) at $\omega = \omega_0$ occurs only when the two dissipation factors are equal, and that the part of the quadrature term which is of the first order in frequency, is zero only under this condition. When either a_1 or a_2 is zero, this first order term is equal in magnitude to the balance of the quadrature term, at the half power points.

If a_1 is zero, Z_2 becomes infinite at ω_0, and zero at the half-power points, as the resistive term vanishes for this choice of a_1. For the other extreme, with a_2 zero, the phase angle of the terminal impedance is arc tan $\sqrt{2}$ at the half-power points, while for the equal dissipation case, this angle is arc tan $2\sqrt{2}$.

Similar conditions apply to the value of the shunt-side terminal impedance Y_1, if the values of a_1 and a_2 are interchanged in the above paragraph.

Thus the choice of a low input-side dissipation factor requires the source resistance to be much smaller than the resistance presented by the transformer in the series-input case and much higher in the shunt-input case, while with a low output-side dissipation factor, the impedance presented to the source near the band-edge becomes very low in the series input case and very high in the shunt-input case.

3.1. *Reflection Coefficient*

The variation of terminal impedances is probably best expressed in terms of the reflection coefficient, ρ, but some care needs to be exercised in the choice of reference impedance when the case with unequal dissipation factors is under consideration. When the source resistance is used as reference, as for example when considering the conditions in a line matched to the source and terminated by the transformer with its load, the reflection coefficient is given by the expression,

$$|\rho_s|^2 = \frac{\varepsilon^4 + \left(\dfrac{a_1-a_2}{a_1+a_2}\right)^4}{\varepsilon^4+1} \qquad \ldots\ldots(8)$$

both for the series-side impedance with R as reference, and for the shunt-side admittance with G as reference.

When the function examined is simply the terminal impedance relative to its value of ω_0, it is more convenient to use this value at ω_0 as the impedance of the

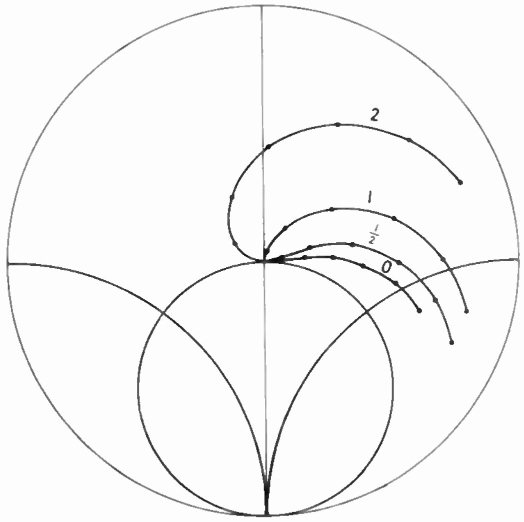

Fig. 3. Reflection coefficient chart.

Series-side impedance with a_2/a_1 as parameter or shunt-side admittance with a_1/a_2 as parameter.
The points marked on the curves are at intervals of 0·2 in ϵ.
The curves are shown for one side of resonance only.

(probably fictitious) reference transmission line. The relative reflection coefficient thus defined is given by

$$|\rho_r|^2 = \frac{\varepsilon^4 + \dfrac{\varepsilon^2}{2}\left(\dfrac{a_1 - a_2}{a_1}\right)^2}{\varepsilon^4 + \dfrac{\varepsilon^2}{2}\left(\dfrac{a_1 - a_2}{a_1}\right)^2 + 4\left\{\dfrac{a_1^2 + a_2^2}{(a_1 + a_2)^2}\right\}^2} \quad \ldots\ldots(9)$$

for the series side impedance or for the shunt-side admittance, in each case with the value at ω_0 as reference.

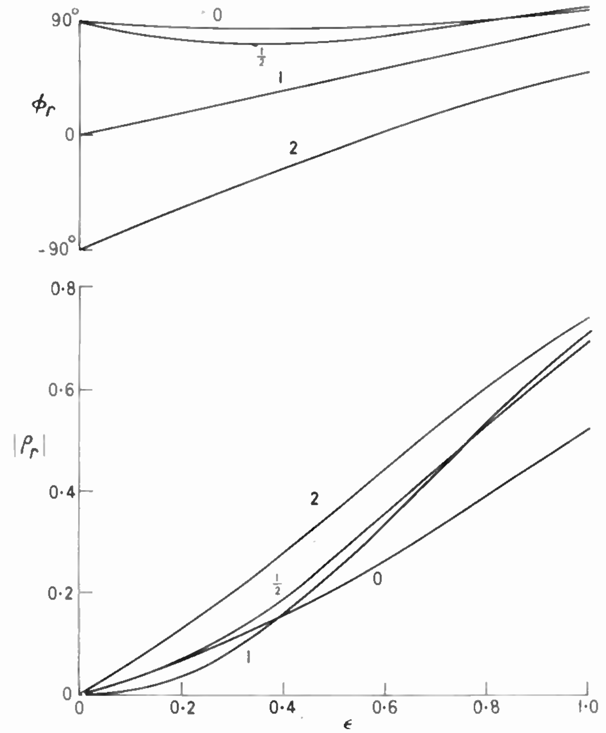

Fig. 4. Reflection coefficient, as a function of the frequency factor ϵ. Parameters as in Fig. 3.

Each of the expressions above reduces to the value given by Stracca, *only* in the equal-dissipation case, giving

$$|\rho|^2 = \frac{\varepsilon^4}{1+\varepsilon^4} \qquad (\text{for } a_1 = a_2)$$

The impedances are presented in the form of a Smith chart plot, using the relative reflection coefficient ρ_r, in Fig. 3; the circles for which either resistive or reactive component is unity or zero are shown for reference. Figure 4 is a graph of the magnitude and phase of ρ_r against ε for several values of the ratio a_2/a_1 for the series-side case. It is evident that for values of ε up to 0·4 (i.e. $|\rho_r| < 0{\cdot}16$), the equal dissipation case gives the smallest reflection, but that further out in the passband the smaller reflections are given by asymmetrical loading with most of the dissipation on the secondary side.

4. Voltage Amplification

When a transformer is used to drive the grid of a vacuum tube from a resistive source, or to couple a resistive load to the anode, the amplification figures of most interest are probably the ratio of grid voltage to the open-circuit voltage of the source in the first case, and the ratio of the load voltage to the grid voltage in the latter, as in Fig. 5.

As the source resistance is considered as part of the network when deriving the transfer impedance, the first of these (Fig. 5(a)) is given simply by

$$\text{amplification} = Z_t/R_s$$

where R_s is the source resistance,

and the latter (Fig. 5(b)) by

$$\text{amplification} = Z_t/g_m$$

where g_m is the transconductance of the tube. Thus in both cases the amplification is just the transfer impedance multiplied by a circuit constant.

From equation (4), it may be seen that when working with a given bandwidth between two fixed resistive loads, maximum amplification is obtained with equal dissipation factors, as the dissipation-dependent part of the transfer impedance reduces to

$$a_1 a_2 (a_1^2 + a_2^2)/a^4$$

which is symmetrical in a_1 and a_2, as would be expected from impedance matching considerations. On

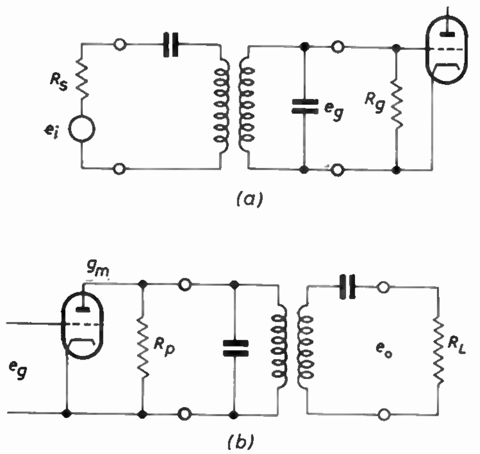

Fig. 5. The transformer as used for grid or anode circuit.

the other hand, when working with given bandwidth between a fixed series-side resistance and fixed shunt-side capacitance as in equation (5), the dissipation-dependent part reduces to

$$a_2(a_1^2+a_2^2)/a^4$$

which has its greatest value in the realizable range when $a_1 = 0$, and $a_2 = a$. This implies that the greatest amplification is obtained when all of the dissipation is in the series-tuned side.[1] For given resistance and capacitance terminations, the voltage amplification so obtainable is twice that obtained with equal dissipations.

5. Design Procedure

In designing a transformer using this method, the value of the factor a has to be chosen from bandwidth considerations, and the grid- or anode-circuit dissipation factor calculated taking into account the grid or anode resistance and all the other stray losses and capacitances associated with the shunt-side circuit. If the value obtained is greater than one half of a, capacitance should be added to bring the factor down to that value for use as an equally-loaded transformer, and if it is less, the greater amplification of the series-loaded transformer may be realized. The losses in the series-side circuit form part of the series resistance R.

When it is desired to connect two valves through a transmission line using two transformers, the equally-loaded design[5, 6] is to be preferred, unless the line is so lossy that the mismatched termination at one end has a negligible effect at the other.

6. Conclusion

This examination shows that although the asymmetrically-loaded transitionally-coupled circuit with all of the dissipation on the series-tuned side and none on the shunt-tuned is capable of twice the voltage amplification of the equal dissipation case, when limited by shunt-side capacitance, it is essential that the resistive source (or load) on the series side remain constant over the passband, and be tolerant of the rather large changes of impedance presented by the transformer if a large fraction of the passband is used.

The series-side resistance can in practice be low enough to match the characteristic impedance of a transmission line, so that the circuit may be used to couple a valve to a resistive load comprising a transmission line with matched termination, or to couple a resistive source through a matched transmission line to the grid of a valve. The method presented here has the advantage that up to twice the voltage amplification of the equally-loaded design may be obtained.

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8. Appendix: Stagger Tuning

The stagger-tuned coupled circuit[7] conditions appeared as an alternative solution in the analysis, and although in general the amplification is lower than with the case discussed above, the results are given here for completeness. (It may be noted that this case coincides with the previously discussed equally loaded case when the two dissipation factors are equal.)

The circuit relationships for this case are

$$L_1 C_1 \omega_0^2 = a_3/a_4 \qquad \ldots\ldots(10)$$

$$L_2 C_2 \omega_0^2(1-k^2) = a_4/a_3 \qquad \ldots\ldots(11)$$

$$\frac{1}{1-k^2} = \frac{a_3}{a_4}\left(2+\frac{a_3^2+a_4^2}{2}-\frac{a_3}{a_4}\right) \qquad \ldots\ldots(12)$$

$$a_3 = \omega_0 L_1 G = R/\omega_0 L_2(1-k^2)$$

$$a_4 = \omega_0 C_2 R = G/\omega_0 C_1$$

$$a = a_3 + a_4$$

$$|Z_t|^2 = 4\frac{a_4^2}{a^4}\frac{k^2}{1-k^2}\frac{R}{G}\Big/(1+\varepsilon^4) \qquad \ldots\ldots(13)$$

$$= 4\frac{a_4}{a^4}\frac{k^2}{1-k^2}\frac{R}{\omega_0 C_1}\Big/(1+\varepsilon^4) \qquad \ldots\ldots(14)$$

where $\varepsilon = \dfrac{\sqrt{2}}{a}\cdot\left(\dfrac{\omega_0}{\omega}-\dfrac{\omega}{\omega_0}\right)$ as previously.

The circuit gives realizable values of k only if

$$a_3^2 > 2\frac{a_3}{a_4} - \frac{4}{1-(a_3/a_4)^2} \qquad \ldots\ldots(15)$$

and

$$a_4^2 > 2\frac{a_4}{a_3} - \frac{4}{1-(a_4/a_3)^2} \qquad \ldots\ldots(16)$$

The series-side terminal impedance, under the conditions discussed in Section 3, is given by

$$Z_t = \frac{R}{a_4} \frac{\dfrac{a_4 k^2}{1-k^2} - \mathrm{j}\left[\left(\dfrac{\omega_0}{\omega} - \dfrac{\omega}{\omega_0}\dfrac{a_3}{a_4}\right)\left\{\left(\dfrac{\omega_0}{\omega} - \dfrac{\omega}{\omega_0}\right)^2 - \dfrac{a_3^2+a_4^2}{2}\right\} + a_3^2\left(\dfrac{\omega_0}{\omega} - \dfrac{\omega}{\omega_0}\dfrac{a_4}{a_3}\right)\right]}{\left(\dfrac{\omega_0}{\omega} - \dfrac{\omega}{\omega_0}\dfrac{a_3}{a_4}\right)^2 + a_3^2} \qquad \ldots\ldots(17)$$

and the shunt-side terminal admittance is given by

$$Y_1 = \frac{G}{a_3} \frac{\dfrac{a_4^2}{a_3}\dfrac{k^2}{1-k^2} - \mathrm{j}\left[\left(\dfrac{\omega_0}{\omega} - \dfrac{\omega}{\omega_0}\dfrac{a_4}{a_3}\right)\left\{\left(\dfrac{\omega_0}{\omega} - \dfrac{\omega}{\omega_0}\right)^2 - \dfrac{a_3^2+a_4^2}{2}\right\} + a_4^2\left(\dfrac{\omega_0}{\omega} - \dfrac{\omega}{\omega_0}\dfrac{a_3}{a_4}\right)\right]}{\left(\dfrac{\omega_0}{\omega} - \dfrac{\omega}{\omega_0}\dfrac{a_4}{a_3}\right)^2 + a_4^2} \qquad \ldots\ldots(18)$$

As shown by Rudd, the shape of the response curve may be made to coincide with the response of the "synchronously tuned" circuit previously discussed, but the maximum voltage amplification will be lower except in the coincident case, and the input impedance will be different.

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The 21st Year of the Parliamentary and Scientific Committee

The end of the 21st year of the Parliamentary and Scientific Committee was marked by a gathering by some of the most eminent public figures in science and parliament at the Annual Luncheon held in London on 15th February, 1962. The former included Sir John Cockcroft, O.M.; Professor A. V. Hill, C.H., F.R.S.; Sir Gordon Sutherland, F.R.S.; Professor Sir Alexander Todd, F.R.S.; Sir Solly Zuckerman, C.B., F.R.S.; and the Institution's President, Admiral of the Fleet the Earl Mountbatten of Burma, K.G.

The guests were welcomed by Dr. Reginald Bennett, M.P., the retiring Chairman of the Committee who, in expressing the Committee's pleasure at the presence of the Chancellor of the Exchequer and the Financial Secretary to the Treasury, said that since communications with the Treasury were currently expressed in verse, he offered the following comment to the Chancellor:

"The Chancellor's concern is great
That we continue to inflate
We sympathise with his desire
That productivity go higher.
If he requires an increment
Of three, or four, or five per cent,
Then surely his best bet must be
On Science and Technology—
Where, for a modest speculation
He may obtain our sure salvation."

Dr. Bennett referred briefly to the Committee's sponsorship during the past year of the European Conference† in which parliamentarians from the Continent as well as Canada had taken part. The membership of the Committee now comprised 130 Members of Parliament and 62 Peers, and 122 Scientific Bodies were represented.

He then welcomed the guest of honour, Sir Howard Florey, President of the Royal Society. Referring to the recent report by the Advisory Council on Scientific policy, Sir Howard said:

"Universities and Technical Colleges have been encouraged to turn out a much larger number of scientists and technologists than they did before the war. So successful does this appear to have been that recently an optimistic forecast was issued which is being held to demonstrate that by 1965 this country will have enough scientists and technologists for all its needs and that by 1970 there will be too many for strictly scientific jobs so that the excess can become administrators!"

He continued: "Because it is possible to issue optimistic figures about the *total* number of scientists being produced it should not be supposed that all is well in Universities and Technical Colleges. Attention in recent years has largely been concentrated on training more and more people for their first degrees, but much less attention has been given to the all-important post-graduate training which in many branches of science is absolutely essential for those who will in the future be engaged in research, and in this connection we have to include a consideration of the opportunities for research available to those who teach the young men and women.

"University Research is now the poor relation of the scientific world in Britain. University scientists who continue to make many of the discoveries of which this country is proud have not the facilities which the pursuit of science in the modern world demands, and so their ability to compete in making discoveries is depressed.

"There is strong international competition for our best scientific men, mature as well as young. These men are looking for facilities to do what they are interested in, and it appears that if the facilities are not forthcoming here some are prepared to emigrate.

"Possibly the difficulties I know to exist in Universities can be likened to the sort of arterial disease from which, on statistical grounds, it can be confidently stated that all of us here are suffering. This disease may not produce overt symptoms but a pathologist knows that it needs as little as a centimetre of the coronary arteries to go seriously wrong to produce crippling consequences. University research is the coronary artery system of the scientific world. It will not matter much what we do to the rest of the scientific organizations if the coronary arteries of University research are allowed to thrombose. In contrast to human coronary disease for which little significant can be done at present, therapy for University science is not impossible: much can be done to improve the present position provided inaction is not justified by procedural wrangles."

The luncheon was preceded by the Annual General Meeting of the Committee which was also a lively occasion. Mr. Austen Albu, B.Sc., A.M.I.Mech.E., M.P., was elected Chairman in succession to Dr. Bennett and, after the formal business, discussion followed on the future activities of the Committee. Among the proposals put forward was one from the General Secretary of the Brit.I.R.E., Mr. Graham D. Clifford, who hoped that consideration would be given to the scientific implications in the research world if Great Britain entered the European Common Market. Mr. Clifford also suggested that the Committee should sponsor more Commonwealth Scientific Conferences.

† See *J. Brit.I.R.E.*, **21**, p. 346, April 1961.

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Pulse Counting and Fast Scaling Transistor Circuits

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AND

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Presented at the Symposium on "Electronic Counting Techniques" in London on 26th April 1961.

Summary: The present state of fast transistor scaling circuits is reviewed and typical designs given for: (1) a frequency divider stage for regular pulses; (2) aperiodic binary circuits for both slow and fast (100 Mc/s p.r.f.) operation; (3) modification of a 4 binary scale-of-16 to a decade scaler operating at 10 Mc/s; (4) general purpose bistable plus AND gates circuits for use in pulse data processing applications as an add and subtract binary scaler or alternatively for gating pulse timing trains up to 8 Mc/s repetition rates. The present maximum repetitive pulse counting rates of scaler design is about 200 Mc/s but these rates may be exceeded with the use of faster transistors or tunnel diode circuits.

1. Introduction

The need for counting the number of pulses occurring within a given time interval is of such common occurrence in electronic instrumentation that suitable circuit techniques are now well known and established. However, the use of counting circuits employing fast transistors has permitted much higher repetition rates than was considered possible a few years ago so that it is of interest to review the present state of these circuit developments. There are a very large number of possible circuit configurations for performing these counting functions and to give an exhaustive treatment of the subject would make this paper too long: instead, a few circuits will be chosen to give an idea of the general principles involved and as being typical of present day practice.

The evolution of circuit designs has led to three main categories:

(1) Frequency division involving the counting of regularly recurring pulses, e.g. pulse generator recurrence circuit design.

(2) Counting of pulses during a given time interval, e.g. measurement of an unknown frequency and counting experiments in nuclear research.

(3) Circuits involving the counting of regularly recurring pulses during a given time interval and obtaining the answer in as short a time as possible, e.g. circuits associated with some types of data processing instrumentation.

The circuit designs required for (3) would perform the functions of (1) and (2) but are sometimes unnecessarily complicated so that simpler circuits are usually employed for (1) and (2).

2. Frequency Divider Circuits

Simple periodic circuits may be used in cases where the input frequency is fixed; if this frequency is variable then the aperiodic circuits of group (2) are needed. A typical frequency divider circuit is shown in Fig. 1 and consists of an emitter-coupled *pnp* trigger pair. The action of the circuit is illustrated by the waveform diagram and it should be noted that the circuit does not self oscillate but the transistors need to be triggered into conduction. This mode of operation is desirable so that in the event of a fault condition

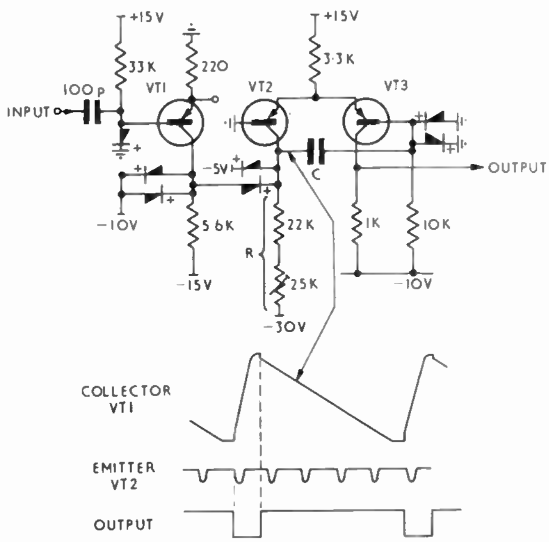

Fig. 1. Pulse recurrence frequency divider.

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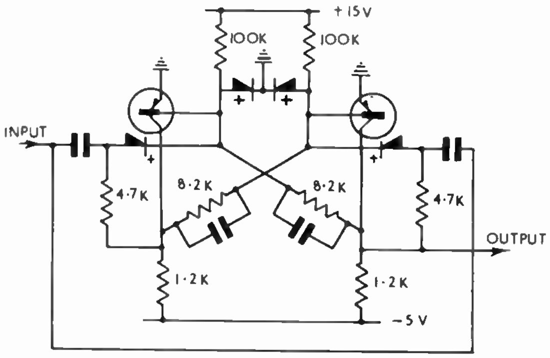
arising (in a chain of such circuits) the final output pulse will probably disappear rather than be present at an altered repetition rate. The amplitude of the trigger pulses passed from one stage to the next are made as small as possible to minimize the effect of any variations of their amplitude on the division rates of the stage. This particular circuit configuration has been chosen to restrict the base to emitter reverse voltage excursion on each transistor to about 0·5 volt thus allowing the use of fast diffused base transistors.

This type of circuit is satisfactory up to input repetition rates in excess of 1 Mc/s and with a frequency division rate up to 10. This division ratio imposes the requirement that R and C (Fig. 1) shall not change by more than ±1% during life.

3. Aperiodic Counting or Scaling Circuits

These circuits are either based on scales-of-two or alternatively ring type scales-of-5 or -10.



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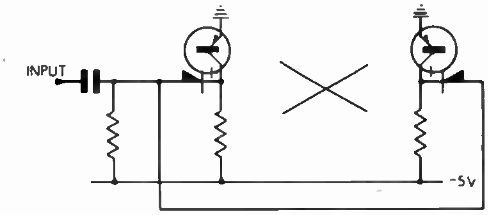

(*b*)

Fig. 2. Saturating binary circuit.

3.1. *Saturating Binary Stage*

Figure 2 shows a common form of circuit where either one or other transistor collector is saturated. This action gives a good definition of voltages and currents in the circuit but results in charge storage in the base region of that transistor which is in conduction. This stored charge must be removed during the triggering action before the circuit can change its state so that the trigger power needed is substantial and the circuit comparatively slow. However, with fast transistors such as type 2N501 a resolution time of 100 nanoseconds can be obtained, permitting counting at 10 Mc/s rates. Also the new "epitaxial" transistors[1] may give improved performance and allow still higher counting rates. The arrangement of the trigger input in Fig. 2 (*a*) is a little more complex than Fig. 2 (*b*) but is superior since it allows the binary collector waveforms to change their state before unduly affecting the relative bias on the trigger "steering" diodes.

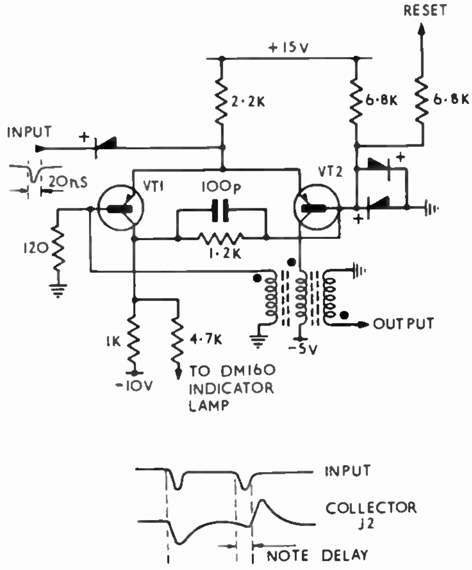

Fig. 3. 10 Mc/s binary (Chaplin circuit).

3.2. *Non-saturating Binary Stage*

Figure 3 shows a design derived from that originated by Chaplin *et al.*[2] The transistor currents are well controlled by the emitter resistor and the values are chosen so that the collector to emitter voltage on each transistor is never less than about 3 volts. Thus the transistor charge storage effects are minimized and the circuit speed is mainly limited by the alpha cut-off frequencies of the transistors. A time resolution of less than 100 ns and counting rates well in excess of 10 Mc/s can be obtained with transistors having alpha cut-off frequencies of 70 Mc/s whilst with higher frequency transistors such as 2N501 a counting rate of 50 Mc/s can be achieved. An examination of the waveform diagram will show that the inductance forms the storage element.

At high counting rates with short duration trigger pulses the effect of the capacitive couplings in the

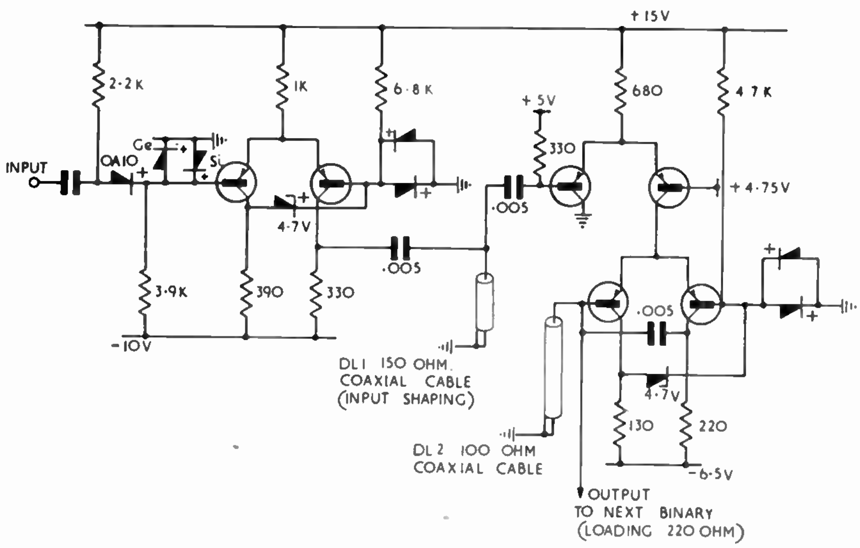

Fig. 4. 100 Mc/s binary with input pulse shaping.

transistors between emitters and bases become serious and leads to a deterioration of the base waveforms. This effect may be minimized by a series transistor in the emitter circuit as in Fig. 4. The capacitive coupling between input trigger pulse and the base waveforms of the binary stage is now reduced to the emitter to collector capacitance of the input transistor. In addition, less power is needed from the input pulses to drive the stage. Figure 4 shows the circuit of a binary scaler with 10 ns resolution time together with an input pulse shaping circuit capable of operating at 100 Mc/s repetition rates whilst Owens *et al.*[3] have achieved counting rates of 200 Mc/s with suitable short duration (2 ns) input pulses. It will be seen that a shorted coaxial pulse shaping cable has been used in Fig. 4 to provide good control of the scaler paralysis and storage time as well as an improvement in waveforms. It would be noted that line DL1 controls the trigger pulse duration to the first binary and this must be rather shorter than DL2 which is the memory element of the binary.

3.3. *Scales-of-Ten*

A decade scaler may be derived from the binary stage design by using four binaries with feedback. One arrangement which has been used at 10 Mc/s is shown in Fig. 5 which illustrates clearly the principles involved. At lower speeds somewhat simpler diode feedback arrangements can be used although the details of circuit design are often somewhat involved. Referring to Fig. 5 the output pulses from the fast binary are normally routed to the second stage except during the period 8 to 10 input pulses when the transistor change-over switch is operated to route the pulse from the first stage to the fourth stage. This changes the scaling factor from 16 to 10. The transistor change-over switch must be timed to operate in between applied pulses and there must be sufficient delay through stages 2 to 4 to allow this timing to be correct.

3.4. *Ring Type and other Circuit Configurations*

Ring type circuits are sometimes used for counting but are especially useful when the circuit has to perform the function of a single pole multi-position switch selecting one out of a number of circuits in sequence. A satisfactory ring-of-ten circuit employs

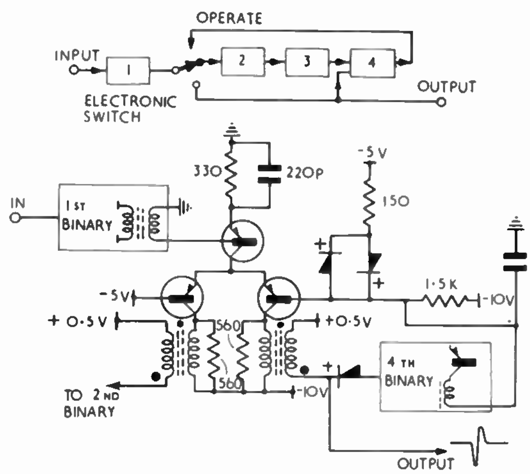

Fig. 5. Feedback arrangement of 10 Mc/s decade scaler.

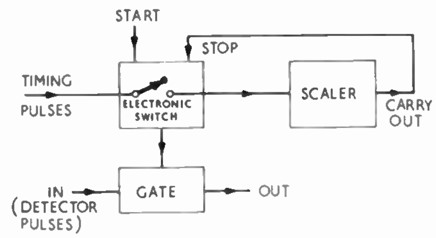

Fig. 6. Precision gate.

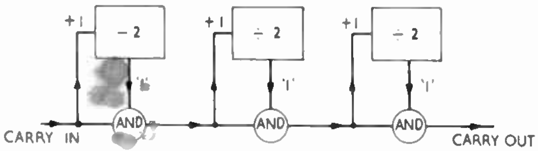

Fig. 7. Fast carry scaler.

ten bistable elements and the *p–n–p–n* device has been found useful for this application at repetition rates up to 100 kc/s.[4, 5] Many other types of circuit may also be used, most of these designs being produced to give low-speed circuits at a minimum cost.

4. Aperiodic Counting Circuits as Logic Elements in Data Processing

In many forms of digital pulse data recording and processing it is necessary to have a general purpose bistable circuit which can be interconnected to form the following main functions:

(1) Add or subtract binary scaler.

(2) A bistable opening a gate circuit.

(3) Shift register.

For some of the scaler applications the time delay through the chain must be kept to a minimum; Fig. 6 shows a typical application in nuclear physics research. In this diagram a pulse gate is opened for a precise number of input timing intervals so that pulses from a nuclear detector may be selected during this gate period. Figure 7 gives the general logic diagram of the system developed for these purposes and the arrangement shows a counting chain. The principle is that the state of the bistables controls the AND gates and allows the input pulses to trigger the appropriate bistable circuits; the particular arrangements will depend on whether an ADD or SUBTRACT scaling action is needed or perhaps shift register operation. Thus the most useful general circuit element for this class of work is a bistable circuit controlling an AND gate.[6] Figure 8 shows such a circuit configuration of a bistable VT3, VT4 controlling two AND gates VT1 and VT2 arranged to give an ADD or SUBTRACT binary scaler operation. The mode of operation may be understood from the waveform diagram but particular points of interest are:

(a) The time delay between input pulses and the carry output pulses is very small, being the transit time of only one transistor.

(b) The carry output pulse is almost the same amplitude and shape as the input pulse. When the binary is in the "1" state the collector of VT3 is at $-4{\cdot}4$ V and the trigger power to drive the binary from the 1 to the 0 state is transmitted from the ADD input pulse through the emitter-base junction of VT1 and then through C2, finally to turn off VT4. The base current in VT1 during this process is sufficiently high to allow collector current to flow to the next stage. VT1 saturates during the ADD pulse and charge is stored in this transistor; this stored charge flows out of the emitter and collector junctions when the base is taken positive to the emitter by the collector of VT3 at the end of the binary change-over period serving to regenerate the

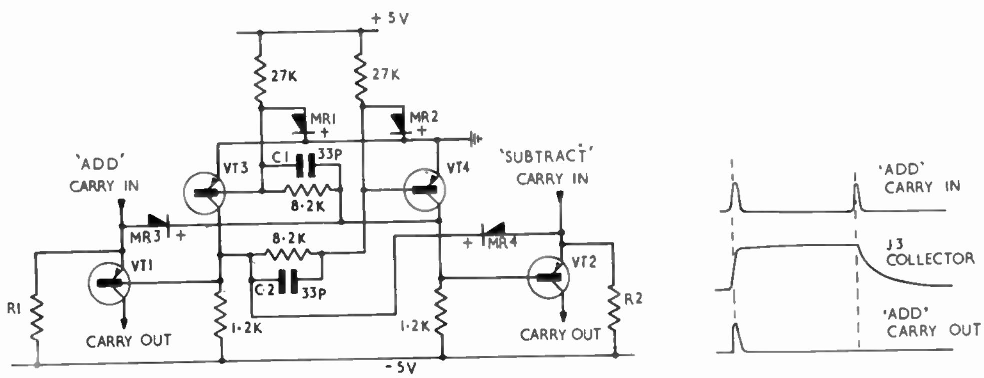

Fig. 8. 8 Mc/s saturated collector bistable circuit with fast add/subtract carry.

pulse slightly. The net result is that about 8 binary stages can be cascaded before the carry pulse is diminished sufficiently to need amplification and reshaping before being applied to the 9th stage.

(c) On the trailing edge of the pulse the charge stored in the stray capacitances and at the collector of the previous transistor must leak away fast enough to allow VT1 emitter to fall faster than VT4 collector. This ensures that VT1 remains cut off when the circuit changes from 0 to 1 and R1 and R2 are chosen to give this condition.

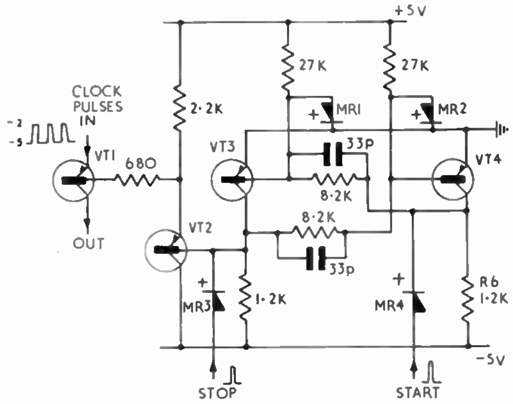

Fig. 9. Clock pulse gate.

Figure 9 shows the same circuit element of Fig. 7 but rearranged slightly to form a bistable opening a gate in a timing chain during the interval between start and stop pulses. This circuit has been used with 8 Mc/s recurrence timing pulses whilst the scaler of Fig. 7 can also be used at similar speeds.

5. Conclusion

This paper has shown that transistor circuits are very well adapted to both fast and slow counting circuits, the upper limit of counting rates at the present being about 200 Mc/s, although this may be exceeded by new developments of fast transistors. However, the advent of the tunnel diode used as a simple bistable element may lead to considerable simplification of these circuits and also permit still higher counting rates.[7]

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Higher Education in Great Britain

The Institution was invited to present evidence to the Committee on Higher Education appointed by the Government on 20th December 1960 under the chairmanship of Lord Robbins, Professor of Economics at the University of London. This paper which has been submitted by the Institution, is based on reports prepared by the Education and Training Committee.†

1. The Committee on Higher Education

The Committee's terms of reference are:

> "To review the pattern of full-time higher education in Great Britain and in the light of national needs and resources to advise Her Majesty's Government on what principles its long-term development should be based. In particular, to advise, in the light of these principles, whether there should be any changes in that pattern, whether any new types of institution are desirable and whether any modification should be made in the present arrangements for planning and co-ordinating the development of the various types of institution."

2. Introduction to the Institution's Submission

One of the important functions of any professional institution is to concern itself with educational matters affecting the profession. The policy it develops through its deliberations takes note of all fields of professional activity and expresses the combined viewpoint of those engaged in research, design, development, production, administration, and technical education.

The influence of a professional body on the education of candidates for membership is likely to be considerable. In the case of technical colleges, the influence is direct through its own examination syllabus and the desire for exemption. In universities and colleges of advanced technology the influence is more indirect although still an important factor.

The interest of the British Institution of Radio Engineers in higher education is reflected in the recent reports which have been published on "The Education and Training of the Professional Radio and Electronic Engineer" and "The Practical Training of the Professional Radio and Electronic Engineer."‡

3. Functions of the Various Types of Educational Institutions Concerned with Education leading to Professional Status

3.1. *Universities and Colleges of Advanced Technology*

At this stage it may be assumed that the Universities and Colleges of Advanced Technology (C.A.T.s) are producing now, and will continue to produce, engineers of a sufficiently high academic standard to meet all foreseeable requirements of the professional institution. Although the methods of the Universities and C.A.T.s undoubtedly differ—and probably should differ—this does not affect the Institution's acceptance of their standards. The products of the University and C.A.T. are also different; the holders of Degrees and Diplomas in Technology may have, in the main, a different approach to their work as engineers yet both attain the academic standard required for professional status.

3.2. *Other Colleges of Technology*

The contribution made by all the other colleges of technology and technical colleges to the training of professional engineers covers a wider range. It should not be accepted that they exist merely to produce technicians and the problem of what part they play in professional education needs examination.

A number of these colleges provide courses for the Diploma in Technology or for external degrees in certain subjects. Although in some the atmosphere of the University or C.A.T. is lacking, there is no reason to object to these courses, as diversity and experiment is desirable in a healthy progressive educational system. But these developments must arise naturally to meet local needs: there is always a danger that colleges may attempt to run advanced courses with the bare minimum of students so that high grade lecturers and expensive laboratories are not utilized to the full.

3.3. *Technical Colleges*

A rather larger group of colleges provides courses where there is a local demand either for the Institution's examinations, or for the Higher National Diploma, or for a college award which may be accepted by the Institution after thorough investigation. This kind of provision needs considerable expansion as without it there are inadequate opportunities for a man in the lower range of professional ability to achieve professional status; possibly propaganda is needed to create a greater demand for this kind of course.

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† Approved by the Council for publication on 15th December 1961 (*Report No. 23*).

‡ "The Education and Training of the Professional Radio and Electronic Engineer", *J. Brit.I.R.E.*, **20**, pp. 643–56, September 1960.

"The Practical Training of Professional Radio and Electronic Engineers", *J. Brit.I.R.E.*, **23**, pp. 171–6, March 1962.

The remainder of the colleges providing advanced courses are concerned with the Higher National Certificate, which gives partial exemption from the Institution's requirements. While this route to professional status is likely to be of diminishing importance, the Institution wishes to emphasize its view that it should be kept open as long as the demand exists.

3.4. *The Need for Part-time Training*

In the last group of colleges mentioned above, a large proportion of the work will be of a part-time nature. The Institution considers that intelligent and mature technicians and other people in engineering employment who, for various reasons, have not received full-time advanced education, should be encouraged to try to qualify for professional status; it follows, therefore, that part-time educational facilities should be retained or extended wherever there is a demand. Part-time education for professional status would normally be of the day-release type with supplementary evening study. The suggestion that for more advanced courses this release should be extended to two days per week or converted to block release has many attractions by providing the additional study time necessary to cover a professional course adequately.

4. The Academic Requirements for Professional Engineers

4.1. *The Basic Requirement*

The Institution has laid down in its Graduateship Examination a minimum standard of academic attainment for radio and electronic engineers which is thought to be suitable for the majority of such engineers employed in professional work in industry. It has always been recognized that the majority of professional engineers do not need to use a more advanced academic knowledge in their day-to-day work and that it is unrealistic to demand excessively high academic standards for membership of a professional institution. Nevertheless it is appreciated that a sizeable minority of professional engineers will be concerned with the most rapid advances in applied science and engineering and must therefore have the highest possible academic standard; for many of these even first degrees are not really adequate qualification and postgraduate academic courses, in addition to postgraduate vocational training, will be needed. There is evidence that this is not yet generally accepted, although the last report of the University Grants Committee makes it clear that the need is recognized by that body. This Institution supports the development of postgraduate schemes rather than the extension of first degree courses to four years.

4.2. *Postgraduate Training*

It is envisaged that many of these postgraduate students will not take further degrees but will take courses directed at specialized applications and vocational needs; on the other hand, university research training needs expansion but it is most important that it should be designed to avoid narrow canalization of interests and at the same time must bring out individual initiative in a way which is rarely possible in research and development organizations outside the Universities. It is, however, possible that a satisfactory or even improved research training can be provided by collaboration between research and development organizations and university departments where the better facilities of the former can be allied with the academic interests of the latter. It is thought that the regulations of many Universities already permit a collaboration of this kind and the Institution would like to see such schemes put into operation.

In addition to the direct value of research training, it is probable that the existence of active research in a teaching department has considerable benefit to the quality of teaching and learning; staff engaged in research are more likely to stimulate students' interests; and if undergraduates are encouraged to mix freely with postgraduate students, there can be no doubt that the enthusiasm of research students in particular must be communicated to the undergraduates.

5. Grants for Postgraduate Students

The Institution views with dismay the continued low level of maintenance grants for postgraduate students. It is thought that a grant which is greatly below a typical starting salary in industry is a positive discouragement to the expansion of research training.

6. Industry's Contribution to Training

The Institution has laid down its policy in regard to practical training in industry. Candidates for corporate membership should desirably have had some form of practical training which should preferably be of postgraduate or sandwich type. The provision of such training is the responsibility of industry which in this context includes Government Departments where appropriate, and the Institution has noted with approval the growth of these facilities and the seriousness with which the training is undertaken.

6.1. *The College-based Student*

The growing problem is the provision of practical training for college-based students: whilst the best means of increasing the number of Dip.Tech. students is by increasing the proportion of college-based students, firms and organizations are finding it difficult to provide practical training other than for their own students.

6.2. *Co-ordination of Academic and Practical Training*

It is felt that there is room for improvement in the co-ordination of practical and academic training and it is in this respect that most progress is required. Too often both in sandwich and postgraduate training, no attempt is made to show the student where the academic processes are involved in engineering design and development.

The growth of both types of sandwich training, thick and thin, is viewed with favour and it is thought that pre-university training is beneficial, particularly in the case of the young student.

Vacation training, although it may be ineffective if not well organized, has an important part to play in the training of a full-time student of engineering, and organizations should be encouraged to continue to provide these facilities.

All industrial training involves co-operation between the college or university on the one hand and industry on the other, but in the case of the thin sandwich courses it is of paramount importance and provides an admirable opportunity of extending contact between the staff of the educational establishment and industry.

7. Summary of the Institution's Recommendations

1. The contribution to higher education of Colleges of Technology should not be ignored. Where Diploma in Technology and external degree courses are conducted and well supported, such courses should be allowed to continue.
2. Opportunity should be available for persons who did not receive full time higher education to achieve professional status by part time means. Thus evening classes and day release classes should be retained and extended if the demand arises.
3. The Higher National Certificate and endorsement certificates for professional status provides a route of slowly diminishing importance to engineering Institutions. However, for the advanced courses, block release or release for two days per week should be considered.
4. The Institution has laid down in its Graduateship Examination a minimum standard of academic attainment for radio and electronic engineers. It is unrealistic to demand excessively high academic standards for membership of a professional institution, although a sizeable minority of professional engineers will require more than a first degree.
5. It is recommended that postgraduate training schemes should be developed and the first degree course limited to a maximum of three years duration.
6. To provide improved postgraduate research training collaboration between research and development organizations and university departments should be encouraged and university regulations amended where necessary to provide for this combination of interests.
7. Existence of active research in teaching departments in Universities and Colleges is of considerable benefit to the quality of teaching and learning.
8. Postgraduate grants must be increased to encourage more graduates to undertake postgraduate research training and academic courses.
9. Sandwich-type training has many advantages and every effort must be made to increase the number of places on "thick" sandwich schemes for university students and on "thin" sandwich schemes for Diploma in Technology students.
10. Industry must be encouraged to make more practical training places available for their own and college-based students.

Electronics at the 1962 Physical Society Exhibition—2

NEW MEASUREMENT TECHNIQUES

A convenient method of measuring time delays in physical systems is by comparison with a concurrent electronic delay. The velocity of sound in matter may be measured in this way, and a method developed at the National Physical Laboratory is being used by the Atomic Energy Research Establishment to study the effects of radiation damage on the elastic constants of irradiated materials. The first pulse from a pulse interval generator is used to start a short burst of sound through the sample. A second pulse is adjusted to coincide with the arrival of the burst at the detector, as indicated on an oscilloscope. The pulse generator settings show the time delay with readings made to a fraction of a millimicrosecond over intervals of up to 100 microseconds. This is achieved by a combination of digital counting technique and precision variable delay line fitted and digital indicator. Use of a 5 Mc/s crystal allows comparison with a broadcast frequency standard, while 50 Mc/s and 500 Mc/s calibration pips on the oscilloscope enable checks on the linearity of the variable delay line to be carried out.

CATHODE RAY TUBE MEASUREMENTS

The vertical deflection amplifier in a modern oscilloscope is expected to have a near-flat frequency response from d.c. up to tens of megacycles. On split-beam c.r. tubes manufactured to date the γ-deflection has been asymmetric, which is a disadvantage in obtaining sweep at high frequencies and in reducing drift at or near d.c. Electronic Tubes showed examples of a new form of construction in which the electron beam is divided at the final anode and a pair of miniature deflection plates are used to deflect each half of the beam symmetrically.

It is current practice for measurements to be taken direct from the screen of a cathode ray oscillograph with an expected accuracy of about 2%. Obviously the tube must therefore be designed and manufactured to produce a display of at least this order of accuracy. The problem of producing an error-free display is complicated by the use of post deflection accleration systems, which cause the path of the electron beam to curve as it approaches the screen. An experimental equipment was shown by Electronic Tubes which has been built to investigate the causes of pattern distortion and non-linearity of deflection. The spot on the tube under test follows a ruled graticule and is kept on the lines of the graticule by means of a photocell and feedback system. The voltages appearing in the feedback system are due to inaccuracies in the tube display. These voltages are amplified and an enlarged picture of the errors, corresponding to the display errors, can be seen.

An apparatus has been developed at the Royal Radar Establishment to examine phosphor screens of high definition cathode ray tubes and to measure the variation in light output of a small sample. It consists of a scanning microscope, photomultiplier tube, directly coupled amplifier and a slow speed oscilloscope. The phosphor under examination is illuminated by a defocused electron beam producing an evenly illuminated area on the screen of the c.r.t. under test. The scanning microscope produces a ten times enlarged image of the phosphor in the plane of a motor driven rotating disc with a circular aperture of 75 microns diameter offset from the centre of the disc by 4 mm. The aperture traces out a circular track within the image of the illuminated area of phosphor thus sampling the variations in light output due to non-uniformity of the screen around the track.

The photomultiplier collects these light variations and passes them via a directly coupled amplifier to a slow speed oscilloscope where they are displayed as changes in vertical deflection as the spot scans across the oscilloscope c.r.t. in synchronism with the rotating disc. The vertical scale in the oscilloscope is displayed as a linear relative light scale; the horizontal scale is proportional to distance around the sampling path of the phosphor under test and is calibrated by scanning a microscope stage micrometer. The scanning aperture of the microscope rotates once every two seconds producing low frequency information. This enables the bandwidth of the amplifier to be limited to about 200 c/s which reduces the photomultiplier noise to a low level. A switch contact is arranged to close once per revolution of the scanning aperture and is used to synchronize the slow speed oscilloscope.

Another use which may be made of the scanning microscope is the plotting of the line width of a c.r.t. trace. A focused line on the c.r.t. may be scanned by the microscope to produce on the oscilloscope a curve of the intensity distribution across the line. The line width may be measured at any particular amplitude by reference to the calibrated horizontal scale. This method is more satisfactory than the usual method of visually estimating the linewidth with a microscope eyepiece graticule.

SEMICONDUCTOR CIRCUITS

High frequency transistors with cut-off frequencies in the region of 800 Mc/s are now readily available, and their properties of high inherent mutual conductance in association with very low values of self-capacitance suggested some years ago that distributed amplifier techniques are not essential in transistor wide-band amplifiers. Development of these basic circuits has resulted in amplifier designs which have gain-bandwidth products in the region of 1·5 Mc/s per stage. An amplifier stage developed by Department of Electronic Engineering, University College of North Wales, Bangor, has a gain of 10 times with 3 dB down points at 10 kc/s and 150 Mc/s. It was designed for use in a 100 ohm system and is suitable for certain head-amplifier applications in nuclear physics. These amplifier techniques were also used in the design of a sensitive wide-band oscilloscope. The c.r.o. tube had a Y-sensitivity of 2 V/cm with a $Y_1 - Y_2$ capacitance of 3 pF, and two stages of amplification were provided in order to give a sensitivity of 100 mV/cm with an input capacitance of 1 pF. The resulting upper cut-off frequency was 100 Mc/s.

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MAGNETIC RESONANCE MEASUREMENTS

The spin-echo principle, subject of a National Physical Laboratory demonstration, originated in 1950 in the University of California. The phenomenon of "nuclear magnetic resonance" (n.m.r.) was by this time well known, and had been used in the study of interaction processes between the various magnetic nuclei in a complex molecule. It is possible by ordinary n.m.r. techniques to measure the "relaxation times" of these processes, namely, the rate at which signals from excited nuclei decay when the excitation is removed and thus give an indication of the relative motions which occur between the molecules of a crystalline of liquid substance. From this information it may be possible to clarify the mechanism of such phenomena as crystal phase changes, changes of state (particularly melting) and self-diffusion in liquids.

The "spin echo" apparatus gives a more direct way of measuring the relaxation times than is possible with a conventional n.m.r. machine, and has the advantage of great versatility in that it can measure time constants over a range of twenty octaves or more, and can also be made to measure independently the two times T_1, due to coupling between the excited nuclei and the remainder of the material, and T_2, due to mutual interaction between adjacent excited nuclei. There is an important difference between these two processes, in that T_1 involves energy loss from the nuclei, whereas T_2 is concerned only with change of entropy.

The Laboratory hopes to be making measurements of T_1 and T_2 on a range of polymer samples in the next few months and relating them to the more elementary physical properties, such as elastic modulus, mechanical damping and dielectric constant, the "end product" of the work being a better understanding of the factors governing the properties of polymers.

An electron spin resonance spectrometer has been developed for scientific research and measurement by Hilger & Watts. Measurements can be made in Q band (35 kMc/s) or in X band (9400 Mc/s) with the same control-consoles, simply by using the appropriate waveguide console and waveguides, and adjusting the magnetic field strength. Electromagnetic energy is fed from a klystron to the H-plane arm of a hybrid-T junction and is split equally between the two side arms. One arm feeds energy to a matched resonant cavity containing the sample; the other carries components which reflect back a reference wave adjustable in amplitude and phase, balanced so that zero power (or the low power needed to bias the crystal detector) is detected in the E-plane arm of hybrid-T. A small coil mounted on the outside of the thin wall of the cavity modulates the steady magnetic field around the specimen at 100 kc/s. Any absorption or dispersion due to electron-spin resonance in the specimen changes the reflection coefficient of the cavity and so unbalances the hybrid-T bridge; the crystal detects the resulting 100 kc/s signal. The signal passes through a low-noise pre-amplifier and a narrow-band amplifier to a phase-sensitive detector, which is also fed with a reference signal adjustable in phase relative to the modulation. A meter presents the detected signal, a cathode-ray-tube displays it, and a pen recorder records it.

A NEW ANALOGUE STORAGE DEVICE

An investigation has been carried out in the Bio-engineering Laboratory of the National Institute of Medical Research to discover whether it is possible to construct a device which can store the magnitude of an analogue quantity without requiring the continuous expenditure of electrical power and without losing the stored information if the power should be switched off. The nearest equivalent to such a device available at present appears to be the electric motor driving a potentiometer. Such a device is clumsy and could not be used in quantity. A number of schemes have been worked out for achieving this object in miniature form. In one of them the optical density of a filter consisting of a transparent base covered with a conducting film on to which a metal can be plated is changed in a reversible manner by plating and unplating from the film. A light and a semiconductive detector is used to read the density of the film which is a function of the value of the analogue variable to be stored. Such devices, apart from their similarity to the integrative units of the nervous system, are thought to find application in analogue computing, learning machines and control devices. Similar devices can be constructed using the progressive magnetization of a hard magnetic material combined with a magnetic resistive detector.