Wireless World

ELECTRONICS, RADIO, TELEVISION

FEBRUARY 1962

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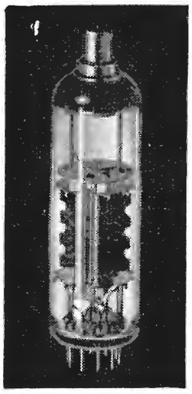


PY800

A New Booster Diode Valve

The PY800 is the new Mullard booster diode which is being used in the latest television receivers. The conditions under which a booster diode is used are such that the valve must be capable of withstanding a high peak inverse voltage and a high peak heater-to-cathode voltage. The P.I.V. rating of the PY800 is 5.25kV and the peak heater-to-cathode voltage rating is 5.75kV. Furthermore, with a heater voltage rating of 19V, it is readily incorporated in a television heater chain. The new booster diode is thus very suitable for timebase circuits using valves with high output powers, and for stabilised timebases.

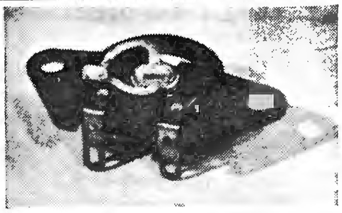
SMALLER TRANSISTORS FOR POCKET RADIOS

With the demand for smaller and smaller portable radios comes the need for smaller transistors. To meet this need, Mullard are now producing miniature transistors less than three-eighths of an inch high and a quarter of an inch in diameter for all the stages of medium and long-wave pocket receivers.

These miniature transistors are encased in a metal envelope from which the leads emerge through a glass button in the base. The leads are so arranged that they can be inserted directly into a printed-wiring grid or can be splayed and connected in the same way as the leads of earlier transistors using the all-glass encapsulation.

The new Mullard r.f. transistors OC44M/OC45M and a.f. transistor OC81M are of this miniature construction, and are being used in present-day pocket sets. The electrical performance of these new transistors is similar to that of the OC44/OC45 and OC81, but their miniature construction enables considerable reduction in the size of receivers (particularly the depth) to be made. The new transistors are thus playing an important role in maintaining a good standard of performance, while contributing to the development of even smaller pocket receivers.

Trimming Potentiometers



Mullard preset resistance controls which can easily be readjusted by the service engineer if the occasion arises, are being used more and more in present-day television sets. These components in the Mullard E097-series are sturdy, small and light, and are extremely efficient and reliable.

The range of nominal resistance offered is from 500Ω to 2MΩ, with ten intermediate values. The minimum resistance of the nominal 500Ω potentiometer is 50Ω, and of the 2MΩ potentiometer is 40kΩ. The variation of resistance with rotation of the slider is linear.

Five versions of the potentiometers are available for various forms of mounting and connection. They can be incorporated in the wiring or screwed to the chassis, or they can be mounted vertically or horizontally on printed-circuit boards. These Mullard carbon trimming potentiometers are thus suitable for use in television sets of any design.

LOCKED-SEAM CATHODES

A new technique for Mullard television valves

All Mullard frame-grid valves for television receivers now use locked-seam cathodes. The normal way of making cathodes is by drawing tubing down to a very small size and then forming the cathode from a cut length of tubing. With the lock-seam cathode, a nickel strip is formed into the required cathode shape in one operation. Experience shows that the latter process results in a much more consistent product which in turn results in valves with more consistent emission and reduced spreads in characteristics.

Watch next month for more information from Mullard about *What's New in the New Sets.*

MVE 241

Wireless World

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The Transistor Symbol

AN ARTICLE on transistor circuit conventions by P. J. Baxandall in last month's issue has revived argument not only on his main theme of circuit arrangement but also because he has expressed a preference, in his introduction, for one particular transistor symbol.

The transistor which made its first impact on the world in 1948 was of the point-contact type, and at once a new symbol was devised which followed closely the physical arrangement of two wire contacts impinging on a flat base of semiconductor material. A period of intensive circuit development followed in which the properties of the point-contact transistor were exploited for amplification and switching. Then came the junction transistor which offered greater power-handling capacity and better constancy of quality in manufacture. Although physically quite different from the point transistor there was no immediate inclination to change the existing symbol and engineers continued to use it as a matter of habit. Perhaps they felt that with new developments coming forward at a rapid pace the junction transistor might itself soon be supplanted and that they would "wait and see" before learning new tricks.

The revolt against the original Bell Telephone Laboratories point-contact symbol began about 1954 and for two principal reasons. Electrically the point-contact and junction transistors differ in many respects, for example in the phase relationship between the emitter and base currents. In some circumstances it was important to know which type of transistor was being used, particularly during the transition period while the junction transistor was establishing itself as the dominant type. The second and more lasting reason was that most users of transistors were already well versed in valve techniques and tended to favour any new symbol which helped them to interpret circuits in terms of valve action. There are pitfalls in carrying this analogy too far, for the current carriers and their distribution in transistors have not the straightforward simplicity of the electron streams in valves, but the occasions on which these differences may lead one into error are few compared with the many cases where the transistor can be safely regarded as a triode valve.

Wireless World supports D. L. A. Barber and W. T. Bane (*Journal of Scientific Instruments*, December, 1956), M. G. Scroggie, P. J. Baxandall, "Cathode Ray" and many others in favouring the symbol (a) or (b) reproduced on page 70 of this issue. We are most of us old enough to have had our grounding in valves (a few, let it be whispered, in sparks and crystals) so we must do a little heart-searching to be sure that our prejudices are not going to hamper the young student and apprentice in his future career. While we believe "our" symbol to be a good one, and hope that the younger generation will appreciate its advantages and widen the sphere of its influence by using it themselves, we must also accept the fact that if they are to go far in their profession they will have to be voracious readers of the world's technical literature. They will find that the point-contact symbol predominates in America and in many Continental countries. Like the notation of music and the shorthand of medical prescriptions it had its origins in the past and has ignored the passage of time. So, too, have some of the many languages of the texts which accompany circuit diagrams in the technical literature of the world. We believe that language is a living thing and should be constantly strengthened by the dropping of obsolete words and refreshed by the acquisition of new. Why should not the same precept be applied in circuitry, as it is in the symbolism of language? And just as the student who hopes to gain mastery of his subject must acquire a working knowledge of more than one language, so must he acquaint himself with more than one transistor symbol. Those who insist that one set of standard symbols is essential to the proper understanding of a circuit do less than justice to the intelligence of the student of today. An unfamiliar symbol may well hold him up while he deduces its meaning from the surrounding circuit context, but the exercise will keep him awake and save him from a too facile acceptance through familiarity. It is better, they say, to chew one's food than to swallow it whole!

That is not to deny the benefits of standardization, but it must not come too soon. We are in favour of keeping an open mind for as long as possible. Although more than a decade has passed since the transistor was invented, new ideas for symbols are still forthcoming. Proposals from the association of French journalists (S.P.R.E.F.) and from other sources are being debated this month at the annual general meeting of the Union Internationale de la Presse Radiotechnique et Electronique in Paris. We do not see why journalists and publishers should not have some say in the matter. After all their drawing offices have to reproduce the symbols and can get through much more work in the day if they have to shift their instruments fewer times.

Until there is world agreement on transistor symbol usage, the policy of this journal will be to keep in circulation all symbols which are easily recognized as transistors.

HIGH-QUALITY TAPE PRE-AMPLIFIER

PREDICTABLE RESPONSE USING HEAVY NEGATIVE FEEDBACK

By P. F. RIDLER,* B.E., A.M.I.E.E.

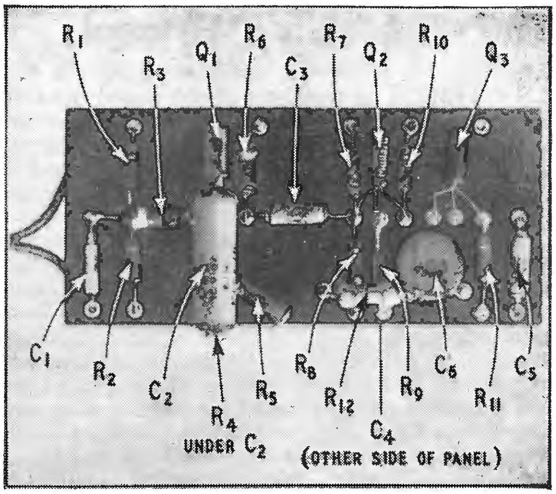
THE essential requirements for a tape pre-amplifier are that it must have a high input impedance, very low noise and predictable response. These are additional to the usual requirements for any audio equipment; namely, that it must be reliable and introduce little distortion.

The requirement of high input impedance can be replaced by that of very low input impedance if suitable heads are available, but as the resistance of most standard heads is too high to give flat response below about 100c/s, the high input impedance amplifier is more versatile [1, 2].

Low noise level is easily achieved using either valves or semiconductors, but is more readily obtained with properly chosen transistors as there is no problem of 50c/s hum to contend with. By using a low noise transistor in the input stage and operating this at a current of about 0.25mA, reasonable gain may be obtained together with a noise level which is quite adequate for the highest quality reproduction.

Predictable response is harder to get with transistors than with valves, but by the use of high feedback ratios the response may be made a function of the feedback network rather than that of the amplifier. Large amounts of feedback involve controlling the gain-frequency response of the amplifier well outside the limits of the frequency band to be covered by the signal input, but with the availability of high frequency transistors this can be done far more easily than a few years ago.

*Salisbury Polytechnic, Southern Rhodesia.



Photograph of tape pre-amplifier showing method of mounting components.

The pre-amplifier of Fig. 1 consists of three stages, the first of which is a low noise transistor (GET106) operated at low collector current. The bias circuit R_1R_2 is isolated from the signal circuit by the 100kΩ resistor R_3 in order that the shunt resistance of the bias potential divider shall not reduce the input resistance; in fact, the input resistance is almost wholly determined by this resistor at high frequencies, where the heavy feedback makes the input resistance of the amplifier alone much greater than 100kΩ. The second and third stages are high frequency transistors (OC44) which are used so that the high frequency response without feedback will be mainly determined by the first stage. This avoids having three high-frequency time constants which would almost certainly cause the amplifier to oscillate at high frequencies when the feedback loop is closed.

The second stage is capacitance coupled to the first and designed for high gain, while the third is an emitter follower to isolate the feedback network from the

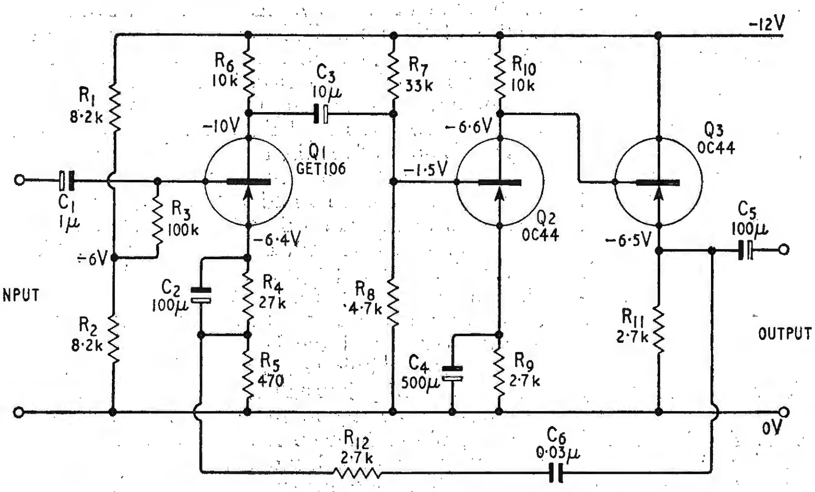


Fig. 1. Circuit of tape pre-amplifier. The voltages were measured with a valve voltmeter. All resistors should be ¼W, ±10% and all capacitors 15V working. R_3, R_5, R_6 and R_{12} should be high stability. For exact values of R_{12} and C_6 see text.

second stage which otherwise would be badly mismatched.

The feedback network has a time constant of 100μsec and reduces the gain by 40dB at high frequencies, providing the rising gain characteristic called for by the C.C.I.R. (7½in/sec) specification at frequencies below about 3,000c/s. The response of the amplifier is shown in Fig. 2, the heavy line being the required characteristic and the points experimental. The feedback also raises the input resistance, this being nearly 100kΩ at frequencies above 3,000c/s. The pre-amplifier is thus suitable for the usual medium impedance head which has an inductance of about 0.5H: if the head has an inductance of 1H the output will be 3dB down at 16kc/s.

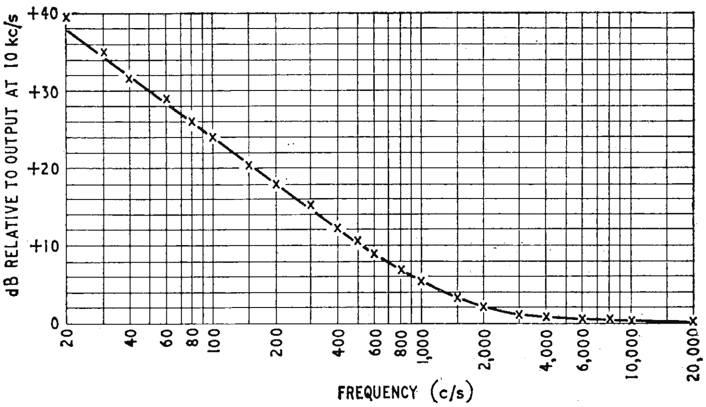


Fig. 2. Response of circuit of Fig. 1. Points are experimental results and heavy line the required characteristic.

The measured noise level over the band 20c/s-20kc/s is greater than 60dB below an input signal of 6mV, which again is typical of the signal produced by available medium impedance heads.

Distortion is unmeasurable, at an output of 300mV, over most of the frequency range, rising to 0.2% at 60c/s (r.m.s. sum of components, major component 2nd harmonic).

The gain is 15 times at 5kc/s so that the average recorded tape (half track) will produce about 100mV, which is enough to drive most amplifiers, fitted with control units, to full output.

Frequency response is within 0.5dB of the C.C.I.R. specification from 30c/s to 20kc/s and is, in fact, flat to about 50kc/s.

The construction is somewhat novel and is illustrated in the photograph. The "chassis" is a piece of $\frac{1}{16}$-in bakelite, and is drilled in the appropriate places and hollow silver-plated rivets (as used in Oak switches) punched in. Components leads are inserted through these rivets and soldered, while inter-connections are made with ordinary wire. This form of construction is easy, cheap to make, and does not present the difficulties of a printed circuit, while retaining its space economy.

The feedback components R_{12} and C_6 will probably be adjusted by trial and error, the easiest way being to use a nominal 0.03μF capacitor and try various resistors until a rise of 10dB is obtained at 530c/s, but if a bridge is available the capacitor can be measured and then a resistor chosen so that $(200+R_{12})\ C_6=10^{-4}$, the 200 being the approximate output resistance of the final stage.

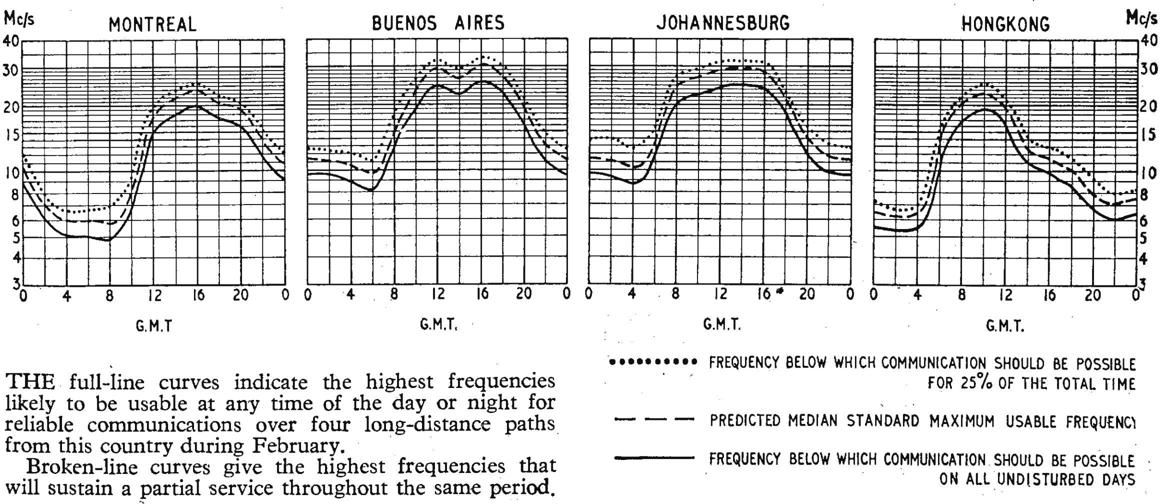
REFERENCES

1. "Transistor Tape Pre-amplifier" by P. F. Ridler, *Wireless World*, December 1958, p. 572.

2. G.E.C. Semiconductor Application Note No. 24.

SHORT-WAVE CONDITIONS

Prediction for February



THE full-line curves indicate the highest frequencies likely to be usable at any time of the day or night for reliable communications over four long-distance paths from this country during February.

Broken-line curves give the highest frequencies that will sustain a partial service throughout the same period.

Computer-aided Air Traffic Control

MINISTRY OF AVIATION EXPERIMENT AT PRESTWICK

AUTOMATIC data-processing equipment (a Ferranti "Apollo" computer) has been installed at Redbrae, near Prestwick airport, where the control centre for trans-oceanic flights is located. It is the first installation of its type in the United Kingdom and it will be used to gain practical experience with computers for a.t.c.

A particular advantage of siting at Redbrae is that all messages concerning the progress of air traffic on the Atlantic routes enter and leave by teleprinter and are already in a coded form suitable for use in the computer.

At the moment, an aircraft crossing the Atlantic first has to clear its flight plan with Redbrae, the controllers viewing the proposed plan and, if necessary changing it so that it does not conflict with other flights. The aircraft then takes off and, maintaining its planned speeds and heights, reports as it crosses each 10° longitude line. Due to unforeseen circumstances the aircraft may have to depart from its planned flight: the controller then notes this in the position report and advises the aircraft if, by deviation from the plan, it is liable to infringe on the internationally-agreed height and along—and cross-track separations, and tells the pilot what action to take to avoid the conflict.

All this has to be worked out quickly and correctly and the situation is further complicated by the vagaries of propagation affecting h.f. R/T communication.

The Ferranti "Apollo" computer is used to perform these calculations for the controller and examine the situation for possible conflicts. To do this it is fed with all relevant flight information: it first compares the proposed flight plan with the situation already present in the machine and routes the result to a teleprinter in front of the appropriate controller (there are four, each looking after a particular section of the traffic). If there is a conflict it advises him and will, when sufficient experience has been gained to programme the machine properly, work out the optimum solution. If the controller endorses the result it is fed back to the originator of the message. Then, as the aircraft's position reports are fed in the computer checks them for deviations and new conflicts.

Computer room. Three left-hand racks contain Apollo: right-hand rack holds conversion equipment for storage tube displays developed by R.R.E.

Controller's position, showing two-colour teleprinter unit typing out from computer in one colour and controller's input in another colour. As soon as appropriate equipment is available automatic strip printers will be used for preparation of flight-progress strips.

A special feature is that the computer will, on demand, present a list of aircraft whose position reports are overdue by a time selected by the controller, so that he may see at a glance whether the likely cause is a failure of radio propagation or something more serious.

At present the machine is working with manual input; that is, all the data to be fed in are first written out and checked for plausibility by an assistant controller. Eventually, however, the computer will do this checking for itself and, as soon as a sufficiently complete programme has been built up, will run on live traffic. For the time being, data comprising past records and additional simulated traffic are being used. Confidence (and programmes) are being built up and already an attempt to feed in incorrect data, such as a five-figure time reference, is detected automatically.

Apollo has, of course, no clairvoyant powers and is in exactly the same difficult position as is a human controller when communication with an aircraft is lost. Steps being taken to improve this are experiments with high-power v.h.f. stations having directional aerials, the possibility of linking extra transmitters by the new submarine telephone cable CANTAT, and by SCOTICE and ICECAN (when they come into service), and the chance of using the met. report teleprinter service (at present transmitted from the U.K. only, on l.f. and which might have its coverage made almost perfect by another station at Gander) for transmission from the ground to aircraft.

Apollo uses transistors throughout, the logic being devised so that among the 350 or so "packages" in the machine only about 20% deviate from six basic types. Most of the computer uses only two types of package—a flip-flop temporary-storage element and a logic gate—and magnetic core stores provide 4,096 words each for programming and data. Initial work indicates that reliability should be good: in the last eleven weeks of operation only three hours have been lost as "down" time. A test calculation is carried out every second and its successful conclusion prevents the ringing of an alarm bell.

WORLD OF WIRELESS

I.E.E. Structure

AS already announced (December p. 616) the scheme for the re-organization of the Institution of Electrical Engineers into three divisions covering electronics, power and science & general, has been approved by the members. The Institution has now issued a list of the Professional Groups within each of the Divisions. There are 28 in all, ten of them being in the Electronics Division which includes within its scope "all matters relating to the generation, transmission, propagation, reception or other use of electromagnetic energy, other than at power frequencies, and components associated therewith whether for communication or other purposes and to the electrical recording and reproduction of sound and vision."

The ten professional groups are:—

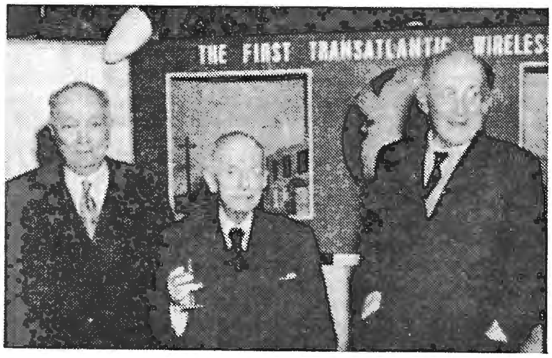
E1 electronic measuring instruments and techniques; E2 computer design; E3 semiconductor devices; E4 components, including valves and tubes; E5 medical electronics; E6 line and radio communication systems; E7 electromagnetic wave propagation; E8 microwave devices and techniques for communication; E9 sound broadcasting and television, including electro-acoustics; E10 radio navigation and radio location.

Student Exchange

SOME 70 delegates from 29 member countries, plus a number of observers, attended the 15th annual conference of the International Association for the Exchange of Students for Technical Experience (I.A.E.S.T.E.) which was held at Imperial College, South Kensington, from January 7th-11th. The president of the conference—the first to be held in London—was Sir Patrick Linstead, rector of Imperial College and president of I.A.E.S.T.E. (U.K.).

Since its formation in 1947 the association has made arrangements for nearly 60,000 students from member countries to receive technical experience abroad during vacations. The subjects of electronics and radio communication are included under the umbrella heading "electrical engineering" in the I.A.E.S.T.E. annual report, but it is obvious from the names of companies participating, both in the U.K. and abroad, that many of the students are in these fields. The 1961 report of I.A.E.S.T.E. (U.K.), of which J. Newby is executive secretary, records that during the 1961 summer vacation 894 students from 48 affiliated universities and colleges were sent abroad for experience (120 of them "electricals") and 965 overseas students came to this country (165 of them "electricals").

Further information on the organization is obtainable from J. Newby, 178 Queen's Gate, London, S.W.7, who was the prime mover in the formation of the association.



Three pioneers of wireless, Capt. H. J. Round, C. S. Franklin and R. D. Bangay, meet again against a backcloth of the first transatlantic wireless stations—Poldhu and St. Johns, Newfoundland—at the Science Museum exhibition of Marconi apparatus. The exhibition, staged to mark the 60th anniversary of the famous "S" signals across the Atlantic, closes on January 25th.

Research in Radio Astronomy.—A Radio Astronomy Planning Committee, under the chairmanship of Lord Fleck, has been set up by the Minister for Science with the following terms of reference: "To consider and advise on the nature and extent of the participation of Government in research in radio astronomy; in particular to consider the programmes, facilities, funds and organization, required to support the U.K. interest in research in radio astronomy both nationally and internationally." The membership of the committee is: —Sir Edward Appleton (Edinburgh University); Dr. D. G. Christopherson (Durham University); Dr. W. L. Francis (D.S.I.R.); Dr. J. S. Hey (Royal Radar Establishment); Professor F. Hoyle (Cambridge University); Sir Willis Jackson (London University); Professor E. R. H. Jones (Oxford University); Sir Bernard Lovell (Manchester University); J. A. Ratcliffe (Radio Research Station, D.S.I.R.); Professor M. Ryle (Cambridge University); and Dr. R. v. d. R. Woolley (Astronomer Royal).

Grants totalling £700,000 towards the cost of building and maintaining two new radio telescopes have been offered by the D.S.I.R. Professor Martin Ryle has been offered £466,000 for a triple parabaloidal radio telescope at Cambridge University, and Sir Bernard Lovell £236,000 for a fully steerable radio telescope at the Jodrell Bank Station of Manchester University.

The Radio Amateurs' Examination, conducted by the City and Guilds of London Institute, 76 Portland Place, London, W.1, is henceforth to be held twice annually. The first examination will be in May, and the second in November. This change is at the request of the Radio Society of Great Britain and the Radio Services Department of the G.P.O.

A.P.A.E. 1962 Exhibition.—Venue of this year's exhibition of the Association of Public Address Engineers is again the King's Head Hotel, Harrow-on-the-Hill, Middx. The date has been fixed for Wednesday, March 7th, and it is proposed to arrange a similar exhibition to last year.

Electricity in Aircraft Control.—Developments in automatic flight control and in airborne navigation systems are two of the subjects to be discussed at a conference on "The importance of electricity in the control of aircraft," organized jointly by the I.E.E. and Royal Aeronautical Society and to be held in London on February 26th-28th, 1962.

I.R.E. & A.I.E.E. Merger?—Joint announcements have been made by the American Institute of Electrical Engineers and the Institute of Radio Engineers that they are considering "the feasibility and form" of the consolidation of the two organizations. The resolution approved by the two institutes stresses that "the advancement of the theory and practice of electrical and radio engineering and the educational and scientific objectives of both institutes may be better served by merger or consolidation . . . into one organization."

Solid-state Circuits Conference.—Five speakers from the U.K. are included in the 46-paper programme of the International Solid-state Circuits Conference, which is taking place on February 14th, 15th and 16th in Philadelphia, Pa. The conference is sponsored by the I.R.E. and a number of other interested bodies. The U.K. participants are A. R. Boothroyd (Imperial College); A. Cole, G. B. B. Chaplin and P. M. Thompson (Plessey); and N. D. Richards (Mullard). Also taking part will be J. J. Sparkes (British Telecommunications Research) and R. E. Hayes (Plessey).

A writ has been filed in the High Court, states J. D. Burke, of Hornchurch, Essex, accusing Thorn Electrical Industries of "infringing his British Patent 877,450 by manufacturing selling renting using television receivers each having a final anode supply system for its cathode ray tube explicitly and substantially described and claimed in the specification."

I.T.A. Tests.—In order to carry out aerial tests in preparation for the construction of the I.T.A.'s new Channel Islands' station at Fremont, Jersey, which will operate in the same channel as Croydon (9) but with horizontal polarization, the trade tests from the London transmitter have temporarily been discontinued on Tuesday and Thursday mornings. The Authority also announces that in future trade tests from Croydon, Mendlesham, Caradon Hill and Stockland Hill will be on half power on Friday mornings. The Black Hill and Durris stations will be on half power on Monday mornings. These reductions in power are necessary to enable one of the two parallel transmitters at each station to be serviced each week.

I.E.E. Activities.—A conference and associated scientific exhibtion covering components in microwave circuits, arranged by the Electronics and Communications Section of the Institution of Electrical Engineers, will take place at Savoy Place, London, W.C.2, next September 19th, 20th and 21st. It is also announced by the I.E.E. that a second symposium on electronics equipment reliability is to take place on the 18th and 19th of October this year.

"Measurement and Control," the new interpretative monthly journal covering mechanical, electrical and electronic instrumentation and control systems, was published by Iliffe Production Publications in January. Edited by T. E. Ivall, at one time an assistant editor of *Wireless World*, its purpose is to "bring out and into perspective the facts that matter to the engineer who uses control equipment, or believes that control techniques can help him in his everyday work." Annual subscription is £3.

Servicing Certificates.—A mistake was made in the results of the R.T.E.B. Servicing Exam. issued by the City and Guilds of London Institute from which we quoted in our December issue (p. 617). Of the 717 who took the final exam. last May 285 (39.8%) passed, 65 (9%) have to retake the practical test and 367 (51.2%) failed.

Southern Instruments Digital Voltmeter-Counter.—In our reference to this equipment on p. 46 of the January issue, the maximum clock pulse p.r.f. was given as 10 c/s. This should read 10 kc/s.

Farnborough Air Shows.—It has been announced by the Society of British Aircraft Constructors that it will not be holding its annual Farnborough show in 1963. Manufacturing members of the Society find the expenses of two major airshows a year (Paris and Farnborough) to be excessive. Whether or not the shows will be held in future in alternate years remains to be seen—the next Paris show is in 1963. The dates of the next Farnborough Show, which has for some years included an increasingly large electronics exhibition, are September 3rd to 9th, 1962.

Car Radio Licences Exceed ½M Mark.—There are now over 500,000 licences issued by the G.P.O. for sound radio receivers fitted in cars. At the end of November last, sound only licences totalled 3.7M, and combined television and sound licences 11.6M.

Personalities

Brigadier E. I. E. Mozley, M.A., A.M.I.E.E., recently appointed Director of Telecommunications at the War Office in succession to **Major General E. S. Cole,** has been for the past three years Commander of the Royal Signals Planning Wing at Catterick where he was concerned with advanced planning in both the tactical and technical aspects of army communications. After graduating at Pembroke College, Cambridge, with a 1st Class in the mechanical sciences tripos in 1936 he served for three years in Malaya as assistant chief signals officer. During the war he held various signals appointments at Catterick, the War Office, and in 21 Army Group before being posted to Lord Mountbatten's staff in S.E. Asia where he was eventually chief signals officer. From 1950 to 1953 Brigadier Mozley was in the chiefs of staff secretariat in Washington, and from 1955 to 1958 he was at the War Office dealing with research and development of signals equipment. One of Brigadier Mozley's recreations is reading for the Bar.

Brigadier E.I.E. Mozley

T. Kilvington, B.Sc.(Eng.), M.I.E.E., recently became secretary of the technical sub-committee of the Television Advisory Committee in succession to **C. W. Sowton,** who relinquished the post on his appointment as staff engineer in the Overseas Radio Planning & Provision Branch (see January issue, p. 14). Mr. Kilvington, a graduate of University College, London, joined the Post Office as a probationary assistant engineer in 1936.

D. W. Heightman, M.Brit.I.R.E., technical director of Radio Rentals Ltd (which he joined in 1956 as chief engineer) and of its associated company Baird Television Ltd., has relinquished these appointments "in order to engage in other activities." He informs us that he is not in a position to comment on his future plans as "negotiations are incomplete." From 1951 to 1956 he was chief television engineer at the Liverpool works of the English Electric Co., prior to which he was on the board of Denco (Clacton) Ltd., which he formed in 1938. It is announced **A. Bamford,** B.Sc., A.M.I.E.E., has been appointed technical director of Baird Television Ltd. He was until recently with Ultra, which he joined in 1942 after graduating at Birmingham University, and has been technical manager of Ultra Radio and Television Ltd. and also a member of the board since the reorganization of the Ultra group two years ago.

T. A. Davies, O.B.E., Inspector of Wireless Telegraphy in the Post Office since 1948, has retired. He joined the Post Office in 1925. In his position as head of the wireless telegraphy section he has been responsible for control and operation of the ship-shore services; inspection of ship's radio equipment; and the examinations for candidates for the P.M.G.'s Certificate.

Reg Billington, T.D., M.Sc.(Eng.), M.I.E.E., is the new Inspector of Wireless Telegraphy in the G.P.O. After obtaining an honours degree at London University, he worked in industry for two years before joining the Post Office Engineering Department in 1936. For a period during the war he was seconded to the Iraq Government as Chief Engineer (P.T.T.) and returned to the Post Office in 1945. Since 1948 he has been Deputy Inspector of Wireless Telegraphy.

R. Billington

M. J. L. Pulling

M. J. L. Pulling, C.B.E., M.A., M.I.E.E., is appointed to the new post of Assistant Director of Engineering in the B.B.C. He will be responsible under the Director of Engineering, **Sir Harold Bishop,** for the Operations and Maintenance Departments, the Engineering Establishment Department and the Engineering Training Department. **F. C. McLean** remains Deputy Director of Engineering responsible to the Director for the Specialist Departments dealing with research, development and new projects. Mr. Pulling, who is 55 and was educated at Marlborough College and King's College, Cambridge, was with Murphy for a few years before he joined the Engineering Information Department of the B.B.C. in 1934. He was superintendent engineer (recording) from 1941 until 1949 when he became senior superintendent engineer (television). Since 1956 he has been Controller, Television Service Engineering.

Hugh S. Pocock, M.I.E.E., editor of *Wireless World* from 1920 until 1941 when he became managing editor, has relinquished the managing directorship of our publishing company, Iliffe Electrical Publications Ltd., to become chairman. He is a director of Associated Iliffe Press. **W. E. Miller,** M.A., M.Brit.I.R.E., the new managing director of Iliffe Electrical Publications, who is also appointed to the board of Associated Iliffe Press, started his technical journalistic career in 1925 when he joined the staff of *Experimental Wireless* (now *Electronic Technology*). A year later he became technical editor of our sister journal *Wireless & Electrical Trader,* of which he has been managing editor since 1955.

P. G. Moger, who has been with Grundig (Great Britain), for six years, is appointed manager of the company's service department in succession to **D. Smith** who has left the organization. Before joining Grundig Mr. Moger was in the R.A.F., where he was engaged on radar development.

W. M. York, commercial director of E. K. Cole, which he joined in 1932, who has also been a director of Pye Ltd. since the merger of the two companies, has resigned from the Pye board because of a reallocation of the responsibilities of Ekco directors within the British Electronic Industries group of companies. **A. W. Martin,** M.B.E., technical director of E. K. Cole, has been appointed to the Pye board, and assumes overall responsibility for the domestic sound radio and television engineering activities of the Pye group of companies, in addition to directing the engineering resources of the Ekco group. Mr. Martin joined Ekco in 1928 as a research engineer, was appointed chief engineer in 1943, and has been technical director since 1952.

A. C. Robb, M.Eng., Ph.D., A.M.I.E.E., who was appointed technical manager of Belling & Lee a few months ago, now becomes technical director. Dr. Robb obtained his Ph.D. degree from Glasgow University, where he took up a research fellowship working on the design of high-voltage particle accelerators. **J. R. Turrill** has become the sales director.

Wing Cdr. R. G. Little, M.B.E., who has been posted to the R.A.F. Central Fighter Establishment as wing commander in charge of electronics, was with the British Joint Services Mission, Washington, from 1959 until recently. He joined the Service as a boy entrant in 1935. He is 42.

R. E. Norman, M.A. (Cantab.), who joined Ferguson Radio Corporation during the war, and since January 1961 has been chief engineer in charge of all domestic television and sound radio activities of the Thorn Group, becomes a member of the Ferguson board of directors.

OUR AUTHORS

G. Buckley, A.M.I.E.E., who describes in this issue an automatic stop for tape or film recorders, has been with the B.B.C. since 1936. For the past ten years he has been in the Designs Department where his particular concern is sound and, to some extent, vision recording. He was for the previous fifteen years in the Research Department.

C. Stott, B.Sc., Grad.I.E.E., whose article on a transistor resistance-coupled oscillator appears on p. 91, is serving a graduate apprenticeship with English Electric Aviation. He joined the company after graduating from Manchester University where he obtained an honours degree in electrical engineering. He is 27.

George H. Olsen, B.Sc., A.M.Brit.I.R.E., author of the article in this issue on field plotting with resistive paper, is on the staff of the Rutherford College of Technology, Newcastle-upon-Tyne, where he teaches physics and mathematics. After war-time service in the R.A.F. he read for a pure science degree at King's College in the University of Durham, graduating in 1950.

OBITUARY

A. Hoyt Taylor, a pioneer of radar in the United States, died in California on December 13th aged 82. Dr. Taylor was chief of radio research in the U.S. Navy from 1922 until his retirement in 1948. As early as 1922 he conducted a series of experiments which showed the possibility of detecting targets by radio. For his work on the development of radar he received the U.S. Medal for Merit in 1944.

Sigurd F. Varian, who with his late brother Russell, invented the klystron tube, was killed in a recent plane crash in the Pacific. The two brothers founded Varian Associates, of Palo Alto, California, in 1948.

FIELD PLOTTING

RESISTANCE PAPER ANALOGUE

By G. H. OLSEN, B.Sc., A.M.Brit.I.R.E.

THERE is an important range of problems in engineering and physics that depends upon a knowledge of the potential field distribution. Typical examples are problems involving fluid flow, heat flow, and gravitation, electric and magnetic fields. In all cases the equation

Flow = Driving force × Acceptance ...... (1)

is obeyed. The driving force is a function of potential V, and the acceptance depends upon the medium involved and the geometry of the surfaces between which the flow is taking place. Examples of eq. (1) of particular interest to electronic and electrical engineers are:—

Current = e.m.f. × conductance .. .. (2)
Magnetic flux = m.m.f. × permeance .. (3)
Electric charge or flux = voltage × capacitance (4)

The flow is real in the cases of heat, current and fluid, but for electric and magnetic fields it is convenient to invent the concept of flow of flux. Usually the flow is in three dimensions. In the absence of sources or sinks of flow materials between the boundary surfaces, the unifying equation related to all of the problems mentioned is due to Laplace. In its cartesian coordinate form it may be expressed as

$$\frac{\partial^2 V}{\partial x^2} + \frac{\partial^2 V}{\partial y^2} + \frac{\partial^2 V}{\partial z^2} = 0 \quad .. \quad .. \quad .. \quad (5)$$

V is the function of potential, x, y, and z being the conventional coordinate directions. It is this equation which must be satisfied at any point in the field within the boundary surfaces.

In certain fields there is no change in field strength in one direction, e.g. the electric field within a long, straight coaxial cable, or the field associated with a straight conductor in a long slot of uniform cross-section.

If the field in this direction is denoted F_z then

$$F_z = \frac{\partial V}{\partial_z} = b, \text{ (a constant)}$$

$$\therefore \frac{\partial^2 V}{\partial_z^2} = 0$$

Eq. (5) thus reduces to the two-dimensional form

$$\frac{\partial^2 V}{\partial_x^2} + \frac{\partial^2 V}{\partial y^2} = 0 \quad .. \quad .. \quad (6)$$

Now the solution to this equation depends upon the boundary conditions, and is obtainable mathematically only in certain simple cases. For example in Fig. 1, which represents the cross-section through a coaxial cable, the potential of the central conductor, radius R_1, is V_1, and that of the outer conductor, radius R_2, is V_2. If the outer conductor is earthed $V_2 = 0$. By performing the partial differentiation it is easy to show that a solution to eq. (6) is

$$V = k \log_e \frac{r}{R_2} \quad .. \quad .. \quad (7)$$

where $r = \sqrt{x^2 + y^2}$ and k is a constant. When, $r = R_1$, $V = k \log_e \frac{R_1}{R_2} = V_1$; and when $r = R_2$ $V = k \log_e \frac{R_2}{R_2} = k \log_e 1 = 0$.

The equipotential surfaces are therefore cylinders concentric with the conductors. The cross-section will have equipotential lines which are circles, the radii of which can be calculated from eq. (7).

When the geometry of the boundaries becomes more complicated a mathematical solution becomes very difficult or impossible. It is under these circumstances that engineers employ analogues to obtain solutions to their potential-field distribution problems.

The principle underlying the various devices used depends upon the analogy that exists between the various forms taken by eq. (1). For example, current is analogous to magnetic and electric flux; voltage and e.m.f. are analogous to m.m.f.; and conductance is analogous to permeance and capacitance. It is possible, therefore, for a system obeying one of the equations (usually eq. (4) for convenience) to be used as a model to solve problems involving any of the other equations. The model is constructed so as to have a field distribution that is geometrically similar to the field which is to be investigated. A major advantage in using a model obeying eq. (4) is the ease with which it is possible to measure voltage as compared with its analogues, m.m.f. in a magnetic field or potential difference in an electric field.

In this article some aspects of one of the possible models, viz. the resistance paper analogy, are dis-

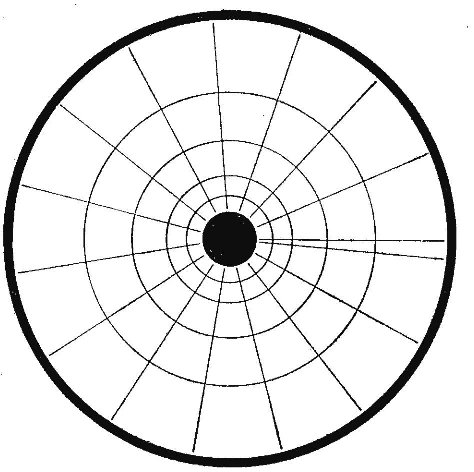


Fig. 1. Field distribution within a coaxial cable.

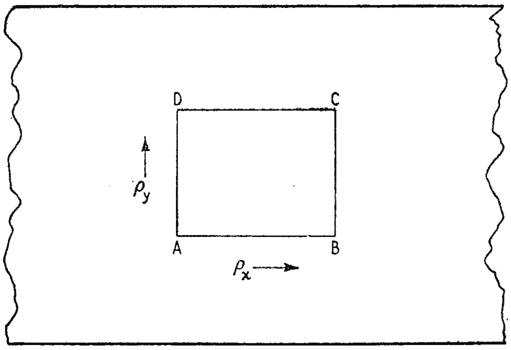


Fig. 2. Finding scale factor for anisotropic resistance paper.

cussed. Such an analogy makes use of the fact that a flow of current through a thin uniform conducting sheet produces a potential distribution that satisfies the two-dimensional form of Laplace's equation. The model is therefore an analogue of a "two-dimensional" field. It is particularly easy to construct and, with care, the inaccuracies need not exceed two or three per cent. A useful paper by J. H. Owen Harries[1] discusses some of the errors involved. The model is particularly useful in schools and colleges where, for teaching purposes, potential distributions can be obtained quite rapidly.

Practical Details:—Telegraph resistance paper is a convenient form of thin conducting sheet. It is readily available[2] as "Teledeltos"* (L39, or equivalent) low-resistance paper, originally developed for use by telegraph engineers for handling ordinary telegrams by facsimile or telephotograph methods. The type used by the writer is available in rolls 50 feet long and 22 inches wide. The resistance between opposite sides of a square was found to be approximately 2,100 ohms in the sample examined. The whole of the lengths of these opposite sides was made conducting in the way described below. (The size of the square is, of course, immaterial; the resistance may therefore be quoted as 2,100 ohms per unit square). Some variation is to be expected from roll to roll; and the resistivity is dependent to some extent on the relative humidity. Provided that the paper is stored in a dry place, however, no trouble ought to be experienced because of local moisture absorption.

A slight disadvantage with this form of conducting sheet is that the resistance of the paper is not quite isotropic. The maximum variation occurs between directions across and along the roll. If the principal axes are taken as being along and across the roll the anisotropic resistance of the paper can be taken into account by applying different scales to the axes. Consider Fig. 2 in which ρ_y represents the resistivity across the roll and ρ_x the resistivity along the roll. ABCD is a rectangle, with AB and AD along the principal axes. Let R_y = the resistance between AB and CD. (For this the lines AB and CD are conductors on the surface of the paper.)

$$\text{Then } R_y = \frac{\rho_y AD}{AB} \quad \ldots \quad (8)$$

$$\text{Similarly } R_x = \frac{\rho_x AB}{AD} \quad \ldots \quad (9)$$

Now ideally when ABCD is a square and

$$\rho_x = \rho_y, R_x = R_y.$$

However when $\rho_x = \rho_y$ the rectangle ABCD on the paper represents a square when $R_x = R_y$, but now AB = AD. To find the necessary scale factor, combining eqs. (8) and (9) gives

$$\frac{AB}{AD} = \sqrt{\frac{\rho_y}{\rho_x}}$$

The factor ρ_y/ρ_x can readily be determined by cutting two squares from the roll in the correct directions and measuring R_y and R_x, when $\rho_y/\rho_x = R_y/R_x$.

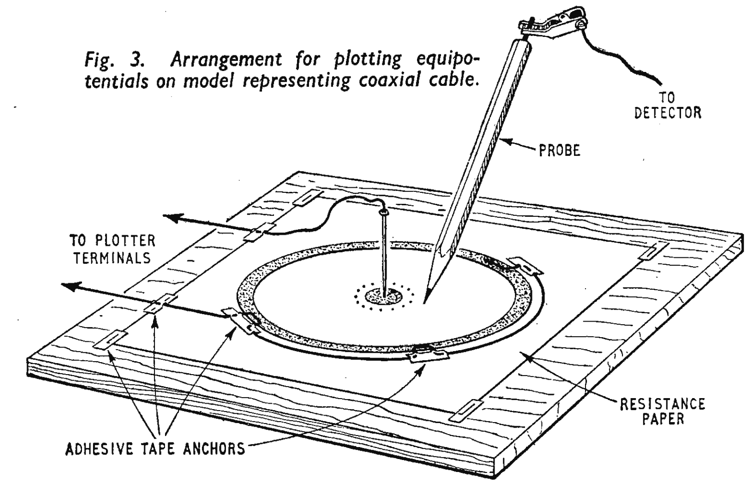
For many purposes the variation of resistance with direction is small enough to be ignored; e.g. when using the analogy to calculate correction terms or to provide no more than the starting figures for a mathematical and exact solution.

The temperature coefficient of this type of paper is low (about 0.2% per °C). Excessive heating of the paper should be avoided by limiting the power dissipated to a maximum of 250 mW per square inch.

The electrodes of the model (corresponding to the boundaries of the real system being simulated) are painted directly on to the paper with silver paint.[3] Care must be taken to stir the paint well before application otherwise the electrode will not be at the same potential throughout. The paper itself is laid on some flat surface. Of the arrangements possible, the writer prefers to use a large drawing board that has previously been varnished, or painted with the type of plastic coating often used as imitation "french polish." The resistance paper can then be held to the board with adhesive tape. The latter may then also be used to anchor the wires making contact with the electrodes. Connections to the electrodes are effected by painting in thin wires so that the latter are embedded in the paint. 28 s.w.g. d.c.c. or enamelled wire, bared at the points of contact with the electrodes, has proved satisfactory.

If only a single contact is made to a long, thin electrode, the latter may not be a true equipotential. The difficulty may be overcome by looping in the

Fig. 3. Arrangement for plotting equipotentials on model representing coaxial cable.



* Standard Telephones and Cables Ltd.

connecting wire at two or three points along the electrode. Fig. 3 gives some idea of a satisfactory arrangement for one model. This method is to be preferred to that of increasing the width of the electrode, since the silver paint is rather costly.

Field Plotting:—Since there are no polarisation effects with graphite resistance paper, d.c. may be used; the plotting is therefore extremely simple.

The apparatus requirements are modest. The essentials are (a) a battery giving 6-12 volts, (b) a calibrated potentiometer, (c) a detector and (d) a probe. The power source will rarely be called upon to supply more than 50mA. The calibrated potentiometer can take the form of a couple of resistance boxes; the total resistance across the supply should be approximately the same as that between the painted electrodes for maximum sensitivity. No serious reduction in sensitivity has been experienced however when using a chain of ten 1% high-stability resistors totalling 5000Ω or when using a ten-turn helical potentiometer of the Colvern type, the total resistance being 10kΩ. The detector used is a 50-0-50μA (1600Ω) meter with protecting resistances and diodes. Fig. 4 shows the circuit used; the battery has been replaced by a full-wave rectifier for convenience. The function of the high-voltage section is explained later. The probe itself may conveniently be a 4H pencil, the graphite core at the top being exposed and connected to the bridge via a miniature crocodile clip and lead. If the electrodes are painted on to the light grey side of the resistance paper the plot of the equipotential lines can be drawn on direct. Some workers have reported a deformation of the field when a particular equipotential is being plotted. However, it is the writer's experience that provided the pencil is kept sharp and the equipotentials merely dotted in at first, and inked or pencilled in after the plot has been completed, no appreciable deformation of the field occurs.

In order to become familiar with the technique the first few plots should be of known fields. Experimental results can then be compared with calculated ones. There is much to commend starting with a plot representing a coaxial cable since the radii of the equipotential lines can easily be calculated from eq. (7). The diameter of the outer painted electrode may be about 8in and that of the inner electrode 1in. Connection to the inner electrode is conveniently made via a 2in brass pin; this then keeps the connecting wire up in the air out of the way. The width of the silver paint representing the outer cable need not exceed one quarter of an inch; but a check must be made to ensure that the electrode is a true equipotential.

After making such a plot with an inner electrode diameter of 1in and the effective outer diameter of 8in the 10% equipotentials were found to be circles. The local variation of paper resistance therefore was negligible for this sample. The following table compares the experimental with the theoretical results. Only in regions of high field intensity (i.e., near the central electrode) were the errors as great as three per cent. Elsewhere the errors were low. It must be remembered, however, that this electrode system can be reproduced very accurately, that the curvilinear squares can be drawn with ease and that errors due to distortion of the current flow near the edges of the paper are eliminated. In most other systems the electrodes must be drawn at the centre of a large expanse of paper to reduce these edge distortions to a minimum.

Equipotential line %	Observed radii (in)	Calculated radii (in)	Error (%)
0	4.00	4.000	0
10	3.25	3.248	—
20	2.63	2.639	—
30	2.14	2.144	—
40	1.75	1.741	—
50	1.42	1.415	—
60	1.12	1.149	3
70	0.92	0.933	1
80	0.75	0.758	1
90	0.60	0.616	3
100	0.50	0.500	0

The field existing between two long, straight parallel conductors (Fig. 5) provides an example for plotting on a large expanse of paper. The shape of the field is well known, so the beginner is readily able to note the effect of edge distortions. The shape of the electrodes painted on the paper, and their distance apart must represent an accurate scale

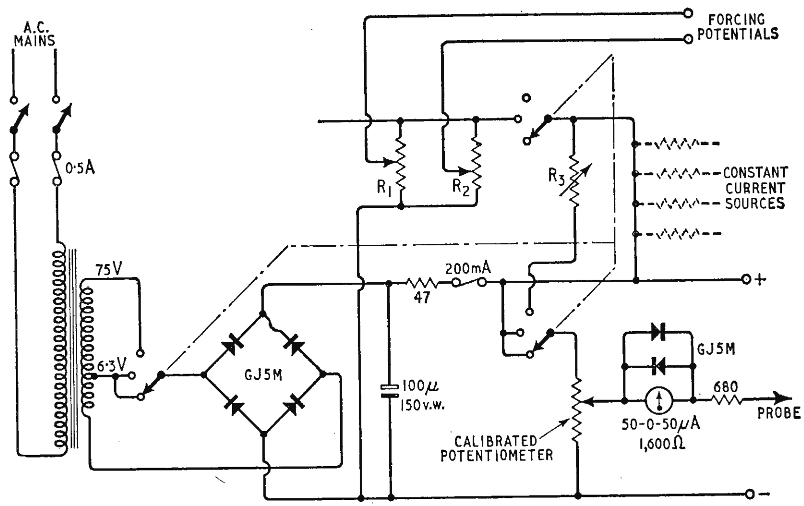


Fig. 4. Circuit diagram of plotter. To obtain plot shown in Fig. 11, $R_1 = R_2 = 2k$. If several forcing potentials are required this value will need to be increased. R_3 depends upon value of constant current resistors. Latter cannot be specified without knowledge of model involved.

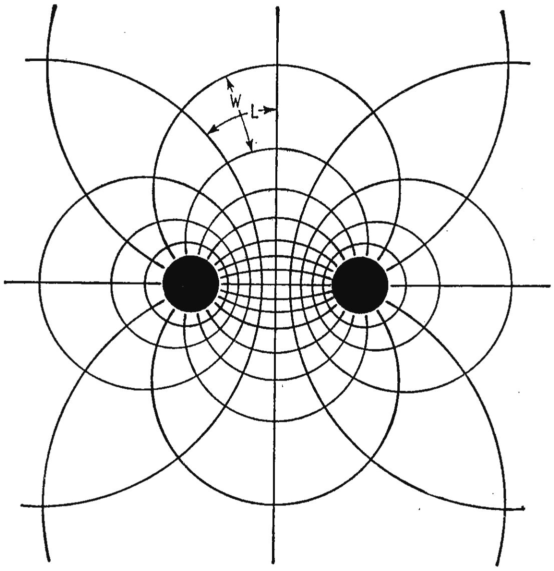


Fig. 5. Electric field between two long straight conductors. Curvilinear squares must have W = L (see text).

drawing of the cross-section of the real system. A suitable scale must be chosen so that the electrodes are surrounded with a large area of paper. Care must be taken to avoid creases when handling these large areas prior to a plot, since the creases have the effect of increasing the local resistance.

After plotting the equipotential lines it is then necessary to sketch in the streamlines representing the lines of flux. All flux lines must cut the equipotential lines orthogonally and must meet an electrode normally. It can be shown that if a set of equipotential lines and flux lines are drawn to form a continuous network of curvilinear squares between the electrode boundaries, then a solution of the Laplace equation has been found. A curvilinear square is an area bounded by four curved lines such that the mean "width" of the area is equal to the mean "length." The tangents to the boundary lines at the corners must intersect at 90°. A test for a curvilinear square is to subdivide the area by equal numbers of intermediate equipotential and flux lines. The smaller curvilinear areas thus created must approach more nearly true rectilinear squares. It will be realized that a particular advantage of plotting and sketching such a network is that, without the use of advanced mathematics, a solution to the Laplace equation can be found which satisfies boundary conditions that frequently are not capable of mathematical expression. Useful information on the sketching of networks of curvilinear squares can be obtained from an article on purely graphical methods of field plotting by E. G. Wright[4] and also in a paper by E. O. Willoughby.[5]

The sketching of curvilinear squares is a matter for trial and error, but for most fields, with the aid of a pair of dividers, it is relatively easy to judge equality of "width" between two equipotential lines (known to be in the correct position) and the "length" between one flux line and the proposed position of the next. A piece of thin transparent sheet with a 90° corner assists in positioning the flux line at right-angles to the equipotential line. The eye should be vertically above the paper when this estimation is made. The edges of the square must be tangential to the flux and equipotential lines at the crossover point.

With fields where it is difficult to sketch the curvilinear squares, trial and error can be almost eliminated by adopting the following method. After painting and energizing the electrodes, sufficient points only are marked along the equipotential lines, no continuous lines being drawn at this stage. It is then usually quite easy to select and sketch with accuracy two flux lines so that a line of curvilinear squares could easily be drawn between them. The original electrodes must then be cut away with a sharp razor edge. The selected flux lines are then painted in and form a new electrode system. (The painting must be within the area between the two lines.) A set of equipotential lines for the second system are then obtained in the usual way. Provided the interval between the lines of the second set is correctly chosen, the first and second sets of lines will combine to form a complete orthogonal plot.

Fig. 6 shows a tube of flux in a two-dimensional field. The flux and the equipotential lines are in a plane which is represented by the resistance paper. The potential difference between equipotential lines across all curvilinear squares is the same, thus it is easy to deduce from the network and the scale of the plot the potential gradient (i.e. the field strength) at any point in the field. For curvilinear squares that are not too unlike rectilinear (or true) squares the field strength is approximately inversely proportional to the length of the side of the square. The field strength is thus highest in regions where the curvilinear squares are smallest.

Behind each curvilinear square for a given depth of field, d, there is a portion of the volume of a tube of flux. Equal quantities of flux pass through each volume section, so each square, irrespective of its size, must represent the same amount of acceptance (e.g. conductance, capacitance, permeance, etc., depending upon the nature of the flux). That this is so can be seen by considering Fig. 7. The volume

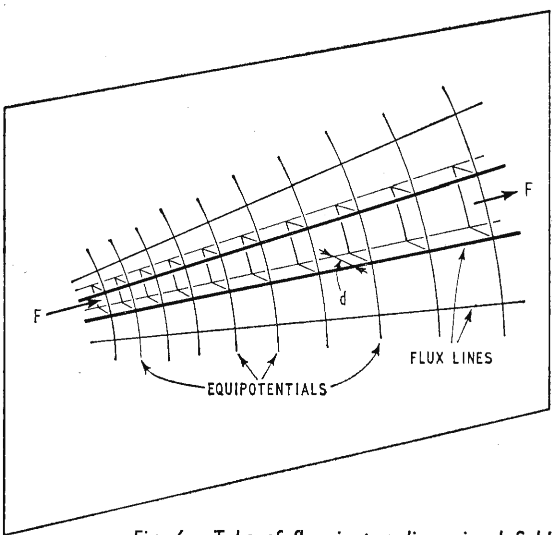


Fig. 6. Tube of flux in two-dimensional field.

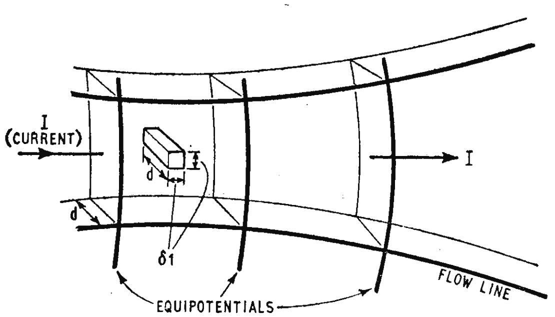


Fig. 7. Determination of conductance.

δv, behind a rectilinear square with length of side δl will have a conductance given by

$$G = \sigma . \frac{\delta l.d}{\delta l} \text{ mho} \quad .. \quad .. \quad .. \quad .. \quad (10)$$

where d is the depth of field. For unit depth of field the conductance of this volume is σ. The volume behind any curvilinear square can be considered as being made up of a large number of volumes similar to δv. Since, by the definition of a curvilinear square, there will be as many δv's along the flux line (i.e. in series) as there are along the equipotential line (i.e. in parallel) the conductance of a volume behind any curvilinear square for unit depth must also be σ. If the plotting of a field between two electrodes yields N_f squares along a flux line and N_e squares along an equipotential line, then the total conductance between the electrodes will be $\sigma N_e/N_f$ mho. (Unit depth of field is assumed in this and following examples.)

Fig. 8 shows the plot representing the current flow in a bus bar of uniform cross-section, made from material of conductivity σ, and pierced by a bolt hole. The current density remote from the hole is assumed uniform. To represent the bar a rectangular piece of resistance paper has a hole cut out to represent the bolt hole, the model being to scale. The conductance of the section shown is $\sigma.12/24 = 0.5\sigma$ mho. The conductance without the hole would be $\sigma.12/21 = 0.57\sigma$mho; so the ratio of the two conductances is 1.14 : 1. It should be noted that N_e need not be an integer. Although N_f need not be integral either, it is usual to plot equipotentials at equal percentage intervals between the electrodes.

By using a similar argument to that for obtaining the conductance of a system, it can be shown that the analogous quantity ' capacitance ' is given by

$$C = \frac{edN_e}{N_f} \text{ Farads} \quad .. \quad .. \quad .. \quad (11)$$

where $e = e_r$ (relative permittivity) $\times\ e_0$ (8.85×10^{-12} F.m.$^{-1}$). For a one centimetre depth of field eq. (11) becomes

$$C = \frac{e_r N_e}{11.3\ N_f} \text{ pF.}$$

Fig. 9 shows the shape of the field existing between two deflector plates of a cathode-ray tube when a given potential difference is applied. The effects of ' fringing ' on the trajectory of the electron, so often ignored when calculating deflection sensitivity, can be estimated. The field strength along the path is determined by comparing the areas of the curvilinear squares. The capacitance of the plates shown in Fig. 9 is 0.6 pF per cm depth of field ($e_r = 1$).

Referring to Fig. 5 the capacitance of the two wires is 15 pF. per metre length when $e_r = 1$. In Fig. 1 the capacitance is $3.5 \times 8.85 \times 15.2/5 = 9.43$ pF per metre length when e_r is taken as 3.5. The calculated value is $24.2 \times 3.5/\log_{10}(4/0.5) = 9.38$ pF per metre length.

The procedure for finding the capacitance of any electrode shape is therefore to paint, to scale, a cross-sectional representation of the electrodes, to plot the equipotential lines, draw in the flux lines to form a network of curvilinear squares, determine the ratio N_e/N_f and apply eq. (11). To determine the capacitance of odd electrode shapes by mathematical means alone would prove difficult or impossible.

A magnetic field in air may be plotted on the assumption that the iron boundaries are magnetic equipotentials. Since iron has a permeability of the order of 1,000 this assumption is usually well justified. Following the lines of eqs. (10) and (11), the permeance of an air gap is given by

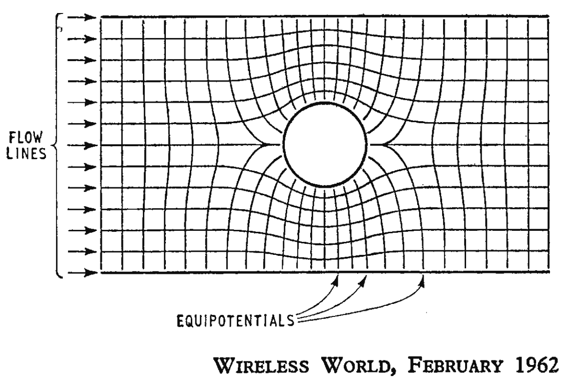
$$P = \mu d \frac{N_e}{N_f} \text{ (webers per ampere turn)} \quad .. \quad (12)$$

where $\mu = \mu_r$ (the relative permeability) $\times\ \mu_0$ (1.26×10^{-6} H.m.$^{-1}$). Frequently in the design of electrical machines the reluctance, S, of an air gap is the more commonly used expression. Since $S = 1/P$ the reluctance is easily calculated from eq. (12). The reluctance of the air gap for the portion of the field AB of an electric machine, shown diagrammatically in Fig. 10, is 220,000 AT/weber. ($d = 1$). The effect of the slots has been ignored; but a further plot may be made to ascertain the reluctance when the slots are taken into account.

A convenient way of plotting flux lines directly is to make use of the fact that a field is completely defined if the positions of either the flux lines or the equipotential lines are known, since one set can be obtained from the other by drawing in a set of curvilinear squares. By painting in electrodes along two known flux lines the resulting ' equipotentials ' for these electrodes will correspond to flux lines in the actual field. In plots used for the design of electric machines, for example (assuming once again a very high permeability for iron as compared with air) on the inverse analogue the portions of the paper representing the iron parts are cut away to form insulators.

The determination of the inductance of current carrying conductors that produce magnetic fields in which there are no iron magnetic circuits may be calculated from a field plot. When currents are fed into the paper via electrodes representing

Fig. 8. Current flow and voltage equipotentials in bus bar with bolt-hole.



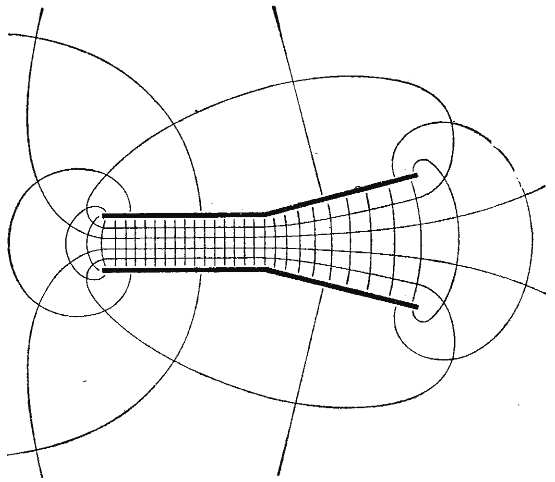


Fig. 9. Field associated with deflector plates of cathode-ray tube.

the real current carrying conductors, the resulting equipotentials on the paper correspond to magnetic flux lines. The analogy assumes all the magnetic flux to link all the current. The contribution of that portion of the field within the conductors linking only part of the current is ignored. Special treatment is required to determine this contribution; but in many cases this is not necessary, e.g. with high frequency currents.

$$\text{Flux } (\Phi) = \frac{\text{m.m.f.}}{\text{reluctance}} = \text{IN} \times \text{permeance}$$

(IN = ampere-turns)

$$\text{Now Inductance (L)} = \frac{\text{flux linkages}}{\text{amperes}} = \frac{\Phi \text{N}}{\text{I}}$$
$$= \text{N}^2 \times \text{permeance.}$$

Using the result of eq. 12 we have

$$\text{L} = \text{N}^2 \, \mu d \frac{\text{N}_e}{\text{N}_f} \text{ henrys} \quad .. \quad (13)$$

In Fig. 1 the central core and outer screen carry currents of I in opposite directions. The resultant m.m.f. outside the cable is zero. Within the cable the radial lines now represent m.m.f. lines and the concentric circles represent magnetic lines of flux. (Note that this is the inverse of the electrostatic case when capacitance was being considered.) The inductance of the cable is

$L = 1.26 \times 10^{-6} \times 5/15.2 = 0.41\ \mu$H per metre length

For Fig. 5 the inductance is $1.26 \times 10^{-6} \times 12/20 = 0.76\ \mu$H per metre length. In both cases N = 1 and $\mu_r = 1$.

The characteristic impedance of a transmission line, Z_0, can be obtained from a single plot by combining the ideas of eqs. (11) and (13). Direct combination is not possible since, although N_f represents the number of squares along a flux line in both cases, this symbol takes a different value depending upon whether the plot represents an electric field or a magnetic field produced by the flow of current in the conductors. However if we allow N_{me} and N_{mf} to represent the numbers of squares along a magnetic equipotential (i.e. m.m.f.) line and magnetic flux line respectively, and N_{ee} and N_{ef} to be the numbers of squares along electric equipotential and flux lines respectively, then $N_{mf} = N_{ee} = N$ (say) and $N_{me} = N_{ef} = M$.

$$\text{For a loss-free line } Z = \sqrt{\frac{L}{C}}$$
$$= \sqrt{\frac{\mu d \, N_{me}/N_{mf}}{ed \, N_{ee}/N_{ef}}}$$
$$= \frac{M}{N} \sqrt{\frac{\mu_0 \mu_r}{e_0 e_r}}$$
$$= 377 \frac{M}{N} \sqrt{\mu_r} \text{ ohms} \quad (14)$$

As an example, suppose a coaxial cable were made from copper conductors the ratio of the screen diameter to the diameter of the central conductor being 8, as in Fig. 1. If the intervening medium had a relative permittivity of 4 the characteristic impedance of the cable would be

$$Z = 377 \frac{5}{15.2} \sqrt{\frac{1}{4}} = 62 \text{ ohms}$$

The paper by E. O. Willoughby[5] gives further examples of the determination of characteristic impedances for various conductor arrangements.

Fields other than electric or magnetic ones can be simulated on resistance paper since the Laplace equation is not confined to these cases alone. A paper by Ross and Qureshi[6] describes the application of the technique to determine the sum of the principal stresses in plates under uniaxial tension when these plates contain discontinuities in the form of cracks. Shear stress distribution is also discussed by them. Streamline flow round aerofoils and heat flow problems may also be undertaken.

In some of these problems it is necessary to

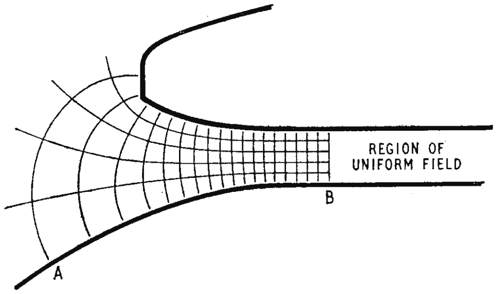


Fig. 10. Portion of field in an electric machine.

have intermediate forcing potentials. For example, Fig. 11 shows the plot representing two pipes, surrounded by insulating material, enclosed in a rectangular box and carrying liquids at different temperatures. To simulate the arrangements electrodes representing the pipes are fed from potential dividers. The voltage between each "pipe" electrode and the electrode representing the outer case may be adjusted with the aid of an Avo 8 or similar meter. This voltage is analogous to temperature difference; the equipotentials represent isothermals and the flux lines correspond to heat flow lines. Fig. 11 is a reduced copy of a plot obtained with a rectangular box 8in. by 6in, the pipes being 2in in diameter. Electrode A was maintained at 5V and electrode B at 3V. This represents the position when the temperature difference between pipe B and the outer case is 60% of that between pipe A and the case. Note that, because of the symmetry of the system, it is not

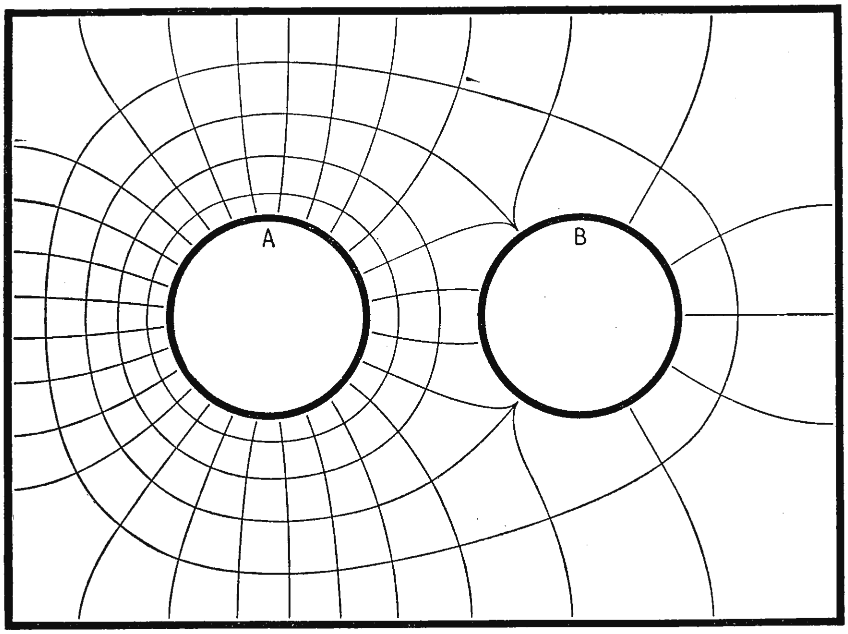


Fig. 11. Field plotting applied to temperature measurements.

necessary to perform the plotting over the whole region. In this case the equipotentials on one side of the straight line passing through the centre of the two electrodes will be identical in shape to those on the other side of the line. In general one should look for lines of symmetry since, if they exist, the work of plotting can be reduced.

In the example just described the electrodes were maintained at constant voltages. In certain cases the electrodes may be required to be fed from a constant current source. Examples arise in the design of electrical machinery where currents fed into the paper give rise to equipotentials that correspond to magnetic lines of flux. To obtain constant current conditions each electrode is fed via a resistor whose resistance is much larger than that of the model. The supply voltage must be increased; hence the presence of the high-voltage section in the power supply. Since the voltages on the paper are much less than the supply voltage it is usually inconvenient to connect the measuring potentiometer directly across the high voltage terminals. To obtain readouts up to 100%, thus allowing more accurate measurements, it is necessary to connect a variable resistor, R_3, as shown in Fig. 4.

It will be seen that the resistance paper analogy is very valuable for obtaining solutions when the worker wishes to avoid lengthy and often complicated mathematical analyses. For those of limited mathematical attainments an insight into many field problems can be gained, when ploughing through the mathematics would obscure the issues involved.

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[1]J. H. O. Harries, "The Rubber Membrane and Resistance Paper Analogies" *Proc. I.R.E.* Feb. 1956 p. 236.

[2]This paper is available from Servomex Controls Ltd., Crowborough, Sussex.

[3]Suitable paint is obtainable from Servomex Controls Ltd., or direct from the manufacturer, Metallurgical Products Ltd. Grade EM50 conducting paint is required.

[4]E. G. Wright, " Graphical Field Plotting " *Electrical Review*, Mar. 1957, p. 507.

[5]E. O. Willoughby, " Some Applications of Field Plotting " *Jour. I.E.E.* Vol. 93 Pt III pp. 275, 1946.

[6]D. S. Ross, and I. H. Qureshi, " The Conducting Paper Analogue and its Application to the Solution of Laplace's and Poisson's Equations." Dept. of Mech. Civil and Chem. Eng., Glasgow Royal College of Science and Technology.

Commercial Literature

Racks, trolleys and multibay assemblies of racking equipment are described in a leaflet from Linvar. Feature of Linvar's rack assemblies is a system of internal baffles that ensures cold air is drawn in for each chassis and hot air is expelled into a "chimney" so avoiding the overheating of equipment " on the upper stories." Linvar Electronic and Mechanical Engineers Limited, Balfour Road, Weybridge, Surrey.

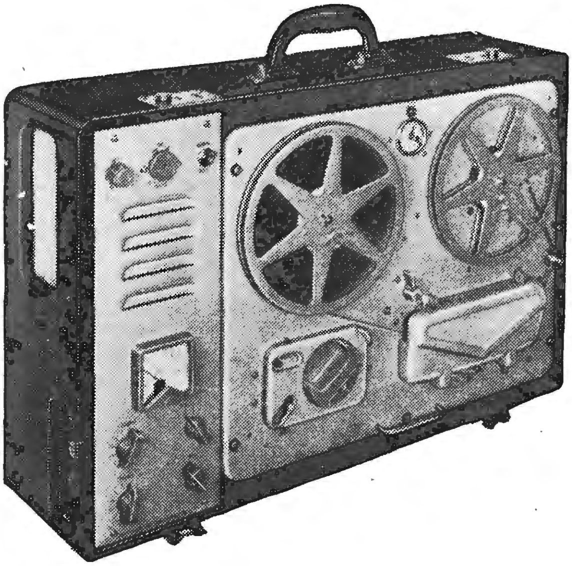
Semiconductors used during development of circuits fail in two ways—either by suffering an electrical catastrophe, or by breakage of their leads due to repeated reconnection. Vero Electronics have designed plug-in holders, which are small boards bearing contacts, to which the devices are attached. Details from Vero Electronics, South Mill Road, Regent's Park, Southampton.

Car-radio aerials made by Valan Electricals include types that can be retracted completely into the wing. Booklet from Valan Electricals, 7 Henshaw Road, Small Heath, Birmingham 10.

Catalogue of, we should imagine, nearly every component that the home constructor (or the professional for that matter) could ever need. Additions in the new Reprint No. 5 include batteries and Continental-type valves. Price 2s. Home Radio (Mitcham) Ltd., 187 London Road, Mitcham, Surrey.

" Printed-Circuit Designer's Handbook ": a guide for the designer as to what can be expected from printed circuits, how to prepare drawings for conversion into printed boards and how to rate conductors and mount components. Copies from Tectonic Industrial Printers Ltd., Denmark Street, Wokingham, Berkshire.

Point plotter, with push buttons for setting in x and y co-ordinates, can be used in conjunction with automatic plotting table for quick production of graphs from data. Four quadrants are covered by negative and positive keys and the plotter may be used as precision d.c. source. Leaflets from Bryans Aeroquipment Limited, 1 & 15, Willow Lane, Mitcham, Surrey.



—Vortexion quality equipment

The W.V.B. recorder has an additional amplifier and head with provision for "before" and "after" record monitoring while the recording is in progress, and this also has echo facilities.

The W.V.A. recorder has provision for a plug in stereo head and can be supplied with this and stereo playback pre-amplifiers with equalisation each having an output of 1 volt from a cathode follower. This is type W.V.A./S.

A heavy mumetal shielded microphone transformer is built in for 15-30 ohms balanced and screened line, and requires only 7 micro-volts approximately to fully load. This is equivalent to 20ft. from a ribbon microphone and the cable may be extended to 440 yds. without appreciable loss.

The 0.5 megohm input is fully loaded by 18 millivolts and is suitable for crystal P.U.'s, microphone or radio inputs.

The playback amplifier may be used as a microphone or gramophone amplifier separately or whilst recording is being made.

The meter fitted for reading signal level will also read bias voltage to enable a level response to be obtained under all circumstances. A control is provided for bias adjustment to compensate low mains or ageing valves.

The power output is 4 watts heavily damped by negative feedback and an oval internal speaker is built in for monitoring purposes.

THE VORTEXION W.V.B. or W.V.A/S

are eminently suitable for making a high quality recording almost indistinguishable from the original since these models have facilities for monitoring the recording actually put on the tape with only a fraction of a second delay.

By this means, when for any reason the signal is distorted or not as required, the result of the recording on the tape can be heard almost instantly, and adjustments can be made until the results are as required.

Many types of music today have the treble boosted considerably, and may result in greater power being recorded at high frequencies than at middle frequencies, an overload of the tape at high frequencies gives a mushy quality with lots of hiss and background noise.

Adjustment to the bias level while listening to the result is useful in this connection especially where the brand of tape and the bias setting for it are not exactly known.

Again if clean treble recordings at $3\frac{3}{4}$in. are of prime importance it is now recognised that no other method is quite so effective in achieving this as reducing the bias slightly while listening to the results. The meter reading of the new bias setting for the particular tape used may be noted for future use.

VORTEXION LIMITED, 257-263 The Broadway, Wimbledon, London, S.W.19

Telephones: LIBerty 2814 and 6242-3. Telegrams: "Vortexion, Wimbledon, Lon."

Transistor Inverters: a Single View

2.—APPLICATION OF GENERAL THEORY

By THOMAS RODDAM

THE first part of this study was devoted to the derivation and examination of the negative resistance characteristic of a transistor circuit, in which positive feedback was provided between the collector and the base. It was shown that this circuit could only rest stably when the transistor was either cut-off or bottomed though it was not explicitly stated that with the inevitable rounding-off which we will have in a practical circuit there will come a critical loading, as the load resistance is made smaller, at which the bottoming will not be very good. We also saw how we can determine which direction the system will move in seeking out one of the two possible stable points.

In the body of the article we also saw how the

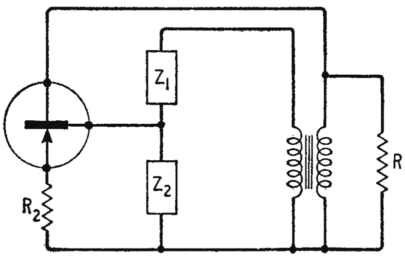


Fig. 13. Basic inverter circuit.

load line or the negative resistance characteristic can be regarded as time-dependent, so that after the system has apparently locked over in one direction it can be dis-engaged so that it must fly over to the other stable point. It was not considered appropriate to point out in the appendix that if the load line lies above the whole of the negative resistance characteristic, which is the dis-engaged condition just mentioned, the whole action is in regions I and IV of Fig. 12 and the trajectory will be followed in the direction of cut-off.

We must now turn to the examination of how this general theory does in fact apply to the inverter circuits with which we are familiar. It will be convenient for reference purposes if we reproduce the basic circuit (Fig. 1 of Part 1) as Fig. 13 and also reproduce the expression for the negative resistance region, which, with the minus sign dropped, is

$$R = \frac{1}{k}\left[\frac{Z_1R_2}{Z_2\alpha} + \frac{R_2}{\alpha} + \frac{Z_1(1-\alpha)}{\alpha}\right]$$

We are free to make this time-dependent by using reactances for either or both of the Z's and we are also free to make the load line time-dependent by using a reactance in the load.

Probably the best known of all the inverter circuits is that shown in Fig. 14. In this circuit we have $Z_1 = R_1$ and $Z_2 \rightarrow \infty$ so that $R = [R_2 + (1-\alpha)R_1]/\alpha$. This circuit is one in which the negative resistance characteristic is fixed and the transition takes place by the rotation of the load line. It is, of course, a push-pull circuit, in which one end or the other is latched except during the transition at the bottomed end, so that we need only think about this end of the characteristic.

The first conclusion to be drawn is that the current must run to the value defined by the top turnover of the negative resistance characteristic in the way already shown in Fig. 7. The second point to note is the fact that if R_2 is at all comparable with $(1-\alpha)R_1$ the negative resistance characteristic will be dependent on R_2 (an internal transistor parameter) and that if it does not depend on R_2 it depends on $(1-\alpha)$, a fairly sensitive quantity which can easily have a range for production transistors of 3 to 1. Consequently the size of the current at which the circuit will unlatch is very dependent on the transistor parameter and in any design for large scale production, when R_1 cannot be trimmed to match the individual transistor, there must be a good big gap between B (Fig. 7), the useful load current, and B″, the unlatching current.

As we have described the circuit the progress towards unlatching along the path B, B′, B″ takes place at almost constant velocity. The current, we say, runs up linearly. If we change the load resistance to some different value we shall move the point B with a proportionate change in the time taken to reach unlatching. A system of this kind is not much good as a generator of alternating current and is really not too good as a d.c. converter, either, because we must arrange our smoothing to deal with the lowest frequency we may encounter.

We come, therefore, to the standard version of

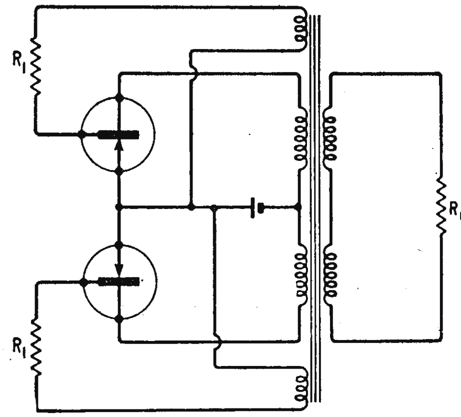


Fig. 14. Classical push-pull circuit.

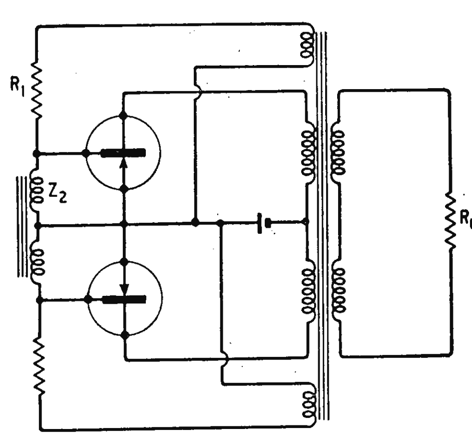


Fig. 15. Using inductive element for Z_2.

this circuit, in which the inductance of the transformer is not constant. I described the behaviour of structures of this kind not too long ago[1] and you may remember that in the idealized form we reached the conclusion that virtually no magnetizing current flowed for a short time, which is inversely proportional to the voltage across the coil, and that then the struggle is abandoned and the current rises very rapidly indeed. If we use a core of this type for the transformer in Fig. 14 we shall not move steadily from B, through B′ to B″ in Fig. 7, but instead we shall remain almost stationary at B for our allocated volt-seconds and then, like Achilles in the fable, make a mad rush for B″. Although we still demand the same peak current, and this is still very dependent on R_2 and on $(1-\alpha)$ the wasted energy is very much less and the timing is almost independent of the load. These properties have been explored in more detail by other writers and you will find the current waveform illustrated in their papers.

In order to eliminate the current spike which precedes the unlocking of the circuit we have just considered we must somehow procure the action shown in Fig. 10, with the load line stationary and the negative resistance moving as time progresses. Two basic methods of doing this are available. We can make Z_1 capacitive or Z_2 inductive. It is usual to make the other element non-reactive. The circuit in Fig. 15 shows the elementary circuit in which Z_2 is inductive. With a normal constant inductance we shall get the constant velocity movement of the point B, B′, B″ in Fig. 14 which we associate with the equation $V = L dI/dt$ and we can expect to see again a serious dependence on transistor characteristics, because of the terms R_2 and $R_1(1-\alpha)$. Here again we can make the inductor with a saturable core so that it dwells at the point B until it has, as it were, run out of volts-seconds, when the negative resistance line will drop sharply to unlatch the circuit. Two practical versions of this arrangement have been published, one by Jensen[2] using a transformer at the base and the other by Berlock and Jefferson[3] using only a simple inverter.

It may be noted that in all these circuits we have discussed so far the term $R_1(1-\alpha)/\alpha$ can be frequency compensated in order to make the negative resistance frequency characteristic an α-dependent one rather than one dependent on α'. The limitation on switching time then becomes partly a problem of the load transformer.

An alternative arrangement is to make Z_1 a capacitor and Z_2 a resistor. This is the circuit shown as Fig. 16 based on Fig. 10, page 296 of the *Mullard Reference Manual of Transistor Circuits*, and is also in essence the classic RC-coupled multivibrator. Here, as in all the linear timing systems, we have the timing dependent on transistor properties so that the frequency will not be particularly well determined. I am not sufficiently up to date on non-linear capacitors to be sure but I do not think that units using the so-called ferro-electric materials are available in sufficiently large capacitance values for use at the audio frequencies which we normally associate with transistor inverters.

It can easily be seen why transistor inverters dislike reactive loads if this method of analysis is followed. We can take the negative resistance line as our reference and cover both forms in a single discussion. The triggering should take place when the load line has moved at a controlled, though not necessarily constant, rate from its first rest point up to the corner where unlatching takes place. With an external reactive load this journey is controlled by the reactance, which is no longer just the shunt inductance we have been considering. When there is some sort of tuned circuit in the load the load line will swing up and down.

There seem to be two factors which must be taken into account here. First of all, with a substantial capacitance the quantity dV/dt which was derived in the appendix to Part I, will be relatively small so that quite an important length of time is occupied traversing the trajectory to the other stable point. Unfortunately, this trajectory will usually pass through a region of high dissipation and we shall get heating problems. The second factor follows from the slow movement along the trajectory combined with the oscillation of the load line. If the load line swings back to intersect the negative resistance characteristic while the working point is still in that part of region III, Fig. 12, in which stability is sought with the system bottomed, there will be a swing back to the original state. Thus a tuned load may give us a complete mode change, probably accompanied by high dissipation,

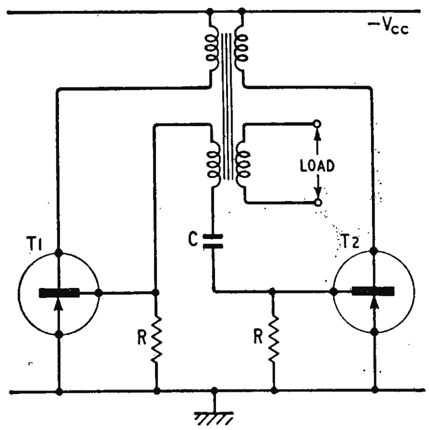


Fig. 16. With Z_1 capacitive and Z_2 resistive, circuit is essentially on RC multivibrator. (Based on Fig. 10 p. 296 Mullard Reference Manual of Transistor Circuits.)

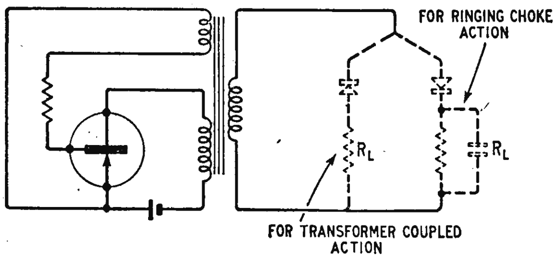


Fig. 17. Two versions of single-ended converter.

or the appearance of part-sinusoidal bites taken out of the square collector waveform. Both these conditions are observed experimentally when saturable elements are used for timing: the saturable inductor continues to measure out its volts-seconds and keeps some hand in what is going on.

Another very important effect in push-pull inverters can be examined rather easily from the basic negative resistance characteristic. When the circuit is unlatched the inductance in the load resistance path, which is the transformer leakage inductance, of course, will try to keep the current constant and will thus send the trajectory off horizontally, parallel to the voltage axis. If the two halves of the collector winding are sufficiently closely-coupled, the latch-in when the second transistor bottoms will catch the collector voltage of the now cut-off transistor. If, however, there is leakage between the two halves of the primary the trajectory can extend well over into a high-voltage area. We are familiar with the appearance of these short spikes at the leading edge and we know that they can cause the failure of our transistor. This behaviour is not immediately obvious when you look at Fig. 12 since the trajectories moving to the right seem to be all below the load resistance line and thus limited to below V_1. The point is that the series inductance will flick the load line up to the horizontal position from which it will then droop, pivoted at the current axis. The reactances in the circuit determine variously how fast the barrier moves down and how fast the trajectory is traced over to the right. They are in a race, the prize being the price of a transistor to you.

De-spiking circuits are frequently used and we can see how they fit into this way of treating the inverter. In the push-pull inverter the load line immediately after unlatching is horizontal, parallel to the voltage axis, at the full unlatch current. The trajectory target is fixed by the other transistor bottoming and because of hole storage effects the turn-on will in fact be faster than the turn-off. There are several ways of expressing the same result now. We can put a small capacitor across from collector to collector in order to absorb the spike, in order to force the trajectory of the unlatched transistor to have a lower speed, in order to make it have a more convenient slope. We can say that as the load is R_L+L_K (the leakage inductance) with plus meaning connected in series, a circuit consisting of $R_L + C_S$ (the spike stopper) will give a purely resistive load line if $L_K/C_S = R_L^2$. This last way, the power circuit method, is the easiest to work out, because a reduced voltage test will show us that the spike duration is τ and we then simply take $C_S = \tau/R_L$ to get an oversafe value, since we must not forget we have stray capacitances too. I make no apology for abandoning the pure single-method discussion because it must give the same sort of answer and will certainly take more time. That is not engineering.

The push-pull inverter circuits are dominated by the transistor which is bottomed. The single-ended inverters are particularly interesting because this symmetry is lost and the two transitions follow quite different patterns. Single-ended inverters are usually required for use as converters and even if they are not being used simply to provide a change of voltage in a d.c.-d.c. system there will be a catching diode in the circuit to provide some of the effects we must now consider. It is usual to distinguish two circuits, the transformer-coupled circuit and the ringing choke circuit. The basic circuit is the same for both and in practice it is the circuit of Fig. 1 in which we make Z_1 a resistance and let Z_2 be infinite. The difference appears when we add the diode as shown in Fig. 17, in which the upper diode is used for ringing choke action and the lower diode for transformer coupled action.

We must begin by considering the circuit bottomed. Unlatching takes place by the sliding of the working point over the corner which we assume for the moment takes place in the linear transformer mode. This brings us to the point P in both Fig. 18 and Fig. 19. When the diode is connected for

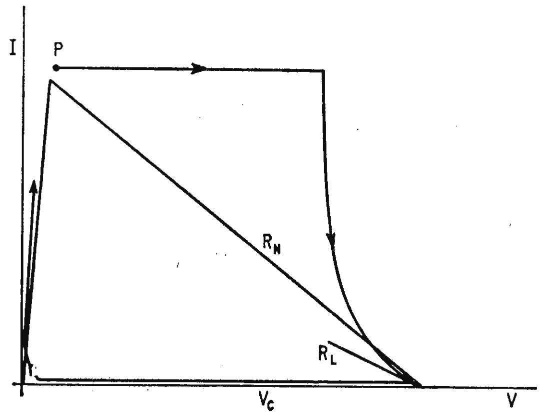


Fig. 18. Working path in ringing-choke action.

transformer-coupled action it will be in the cut-off sense for V increasing: when it is connected in the ringing choke mode we assume a storage capacitor, already charged, which will hold the diode cut-off. We thus have a load line passing through the point P but having a slope corresponding to an open-circuit. The trajectory is therefore roughly horizontal up to some voltage V_c, the voltage to which the storage capacitor was charged, a point reached after a time determined by the velocity along the trajectory which in turn depends on the stray capacitances. As soon as V_c is reached the diode conducts and the load line becomes that of the capacitor, almost a short-circuit. The trajectory is now initially a vertical, constant voltage line, swinging round as the capacitor charge increases its voltage, and becoming asymptotic to the slope R_L. This is shown in Fig. 18.

The transistor is now cut off, and so is the diode. A new load line, again corresponding to infinite impedance, lies close to the I = 0 axis (the effect

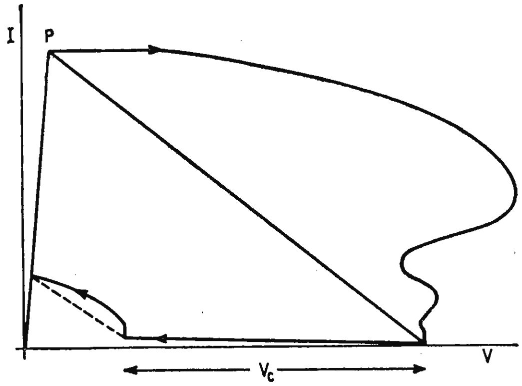


Fig. 19. Working path in transformer-coupled action.

of I_{co} is neglected). The only intersection of this with the R characteristic is right down near the origin so that this point is now sought. We are thus all prepared for the linear run up the so-called diode line to the unlatching point.

Although in Fig. 18 the value chosen for V_c is below the supply voltage there is no reason why it should be. The only real determining factor will be the fact that the energy pumped into the capacitor during the charging period must, in the steady state, equal the energy flowing away through the load resistance. This circuit seeks out that balance and thus has very poor voltage regulation.

If we connect the diode the other way round we do not get the sharp catching action at V_c. The trajectory depends entirely upon the storage and takes the form shown pretty arbitrarily in Fig. 19. Just how high the voltage will swing is anybody's guess and in all practical circuits a second diode, a spill-over diode, is used to catch the trajectory. If good regulation is wanted this diode is not used to deliver power to the load but is used, instead, to deliver power to the battery. The path along the $I = 0$ axis is followed again, provided that there is a storage capacitor, but when the voltage has dropped by an amount equal to V_c the diode begins to conduct. The load line is now, for the moment, vertical and we get the dog-leg shown in the lower left-hand corner of Fig. 19. This phase ends when bottoming is reached.

Since the working action of the transformer coupled circuit takes place below the negative resistance line we are free to allow this line to move. It is thus possible to make use of a saturable core in this sort of circuit. The ringing-choke inverter cannot be operated in this way. The practical problem, which makes this not too popular, is that the flux must be re-set by an auxiliary winding. It will, I hope, be clear to the reader that he could use the CR or RL couplings to produce a single-ended inverter and that in the transformer-coupled form he can omit the storage capacitor if he wants a straightforward pulse generator. He may even find some other circuits which I have omitted which fall into the same broad class.

Obviously this survey of a number of inverter circuits is by no means complete. It was not intended that this should be an article on inverter design. Any design study of a particular circuit must take into account some of the factors which we have omitted. A real transistor has a finite leakage current and does not start to conduct until a small forward bias is provided at the base. Transformer and inductor are not pure reactances and the cost of these components is often closely related to the amount of resistance which can be tolerated. Two circuits which are theoretically the same may, because of these factors, show important practical differences.

From a beginner's point of view, the point of view of a man who wants to know how the circuits work, this treatment seems to have quite considerable advantages. The similarities between the circuits and the really fundamental differences show up clearly. There is a fairly simple mathematical treatment which is common to the whole group and the idea of making allowance for a less idealized system is not really too repugnant; it only needs to be done once, after all.

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[2] "An Improved Square-Wave Oscillator Circuit," by J. L. Jensen, *I.R.E. Trans. on Circuit Theory*, Vol. CT-4, Sept. 1957, p. 276.

[3] "Transistor Inverters and Converters," by M. D. Berlock and H. Jefferson, *Wireless World*, Aug.-Oct. 1960.

Industrial Groups—VII

AT the eighty-seventh annual general meeting of **Gas Purification & Chemical Co. Ltd.**, held recently in London, Douglas D. Mathieson advised his intention to resign the chairmanship, whilst remaining a director. Joseph Green was subsequently elected to that office.

The "radio and electronics" content of the group, whose original business, as its name implies, was (and still is) in the chemical industry, has grown considerably since it acquired Grundig (Great Britain) Ltd. seven years ago. In the past eight years its gross trading surplus grew from £14,000 to a peak of £800,000, but last year there was a trading deficit of some £14,000.

In 1954 B. & R. Relays joined the group, and in 1955 Wolsey Electronics, the aerial and relay equipment manufacturers, were acquired. In the following year Staar Electronics was formed within the group to manufacture, in this country, the automatic record-playing equipment developed by U. G. Staar, of Brussels. A.B. Metal Products, component manufacturers, came into the group in 1957.

Practically the whole of the 1960-61 loss is attributed to those campanies in the group which are concerned with radio, television and tape recorders.

During 1960-61 a number of changes affected subsidiary companies: the name of Greencoat Electronics (formerly Staar Electronics Ltd.) was changed to Greencoat Industries Ltd., and the name of Johnson British Electric Ltd. was changed to Johnson British Wood Crafts Ltd. The following wholly-owned subsidiaries were wound up: G.P.C. Electronics Ltd., Kingsway of London Ltd. and Plastic & Metal Products Ltd.

It was also announced that the sale of the machine tool division of Smart & Brown had been effected for £300,000, and that the electrical side was being transferred to A.B. Metal Products. The group's present constitution is:—

Gas Purification & Chemical Co.	Grundig (Great Britain)
A.B. Metal Products	E. G. Irwin & Partners
B. & R. Relays	Johnson British Wood Crafts
British Industrial Holdings	Modern Machine Tools
Francis Street Properties	Plyglass
Gas Purification	Precision Plastics
Greencoat Industries	Smart & Brown (Connectors)
	Wolsey Electronics

LETTERS TO THE EDITOR

The Editor does not necessarily endorse the opinions expressed by his correspondents

Safety of Life at Sea

WE were very interested in Mr. A. T. Ferguson's letter (December 1961 issue), particularly in regard to the aid to be given to crash survivors.

Economically, it is apparent that the suspension of d.f. facilities from British coast radio stations is justified. Originally, back in the nineteen twenties, the stations were erected primarily to sell bearing information to shipping. A nominal charge of five shillings was charged for a "fix", the resultant revenue being sufficient to defray part of the maintenance costs. However, during the last few years, the use of electronic navigational equipment has become a standard facility in most ships and consequently the service offered by the d.f. stations has been more and more neglected.

Unfortunately, any method must be financed, and the question arises—is the retention of the coastal d.f. facility a major factor in the preservation of life at sea? A review over recent years does not disclose any significant event in which the present d.f. facility alone was responsible for determining the position of survivors.

Currently, survival kits for dinghies and ships' lifeboats contain radio equipment which operates within the frequency band of the coastal d.f. stations. When operated under ideal site conditions, a medium-frequency transmitter has considerable range. Several variable factors determine the range figure but the aerial system is the one that is the most restricted when used in small craft at sea. A quarter-wave aerial for 500 kc/s is obviously prohibited by physical dimensions alone, and furthermore the alternative method of lumping the inductance to tune a shorter aerial results in decreased efficiency of the radiating system. In practice, a lifeboat carries a 30ft. length of wire which can be rigged as an aerial and, using such an arrangement, a range figure of up to approximately 100 miles can be achieved.

Results are much worse when the equipment is operated from a rubber dinghy. Under these conditions, a whip aerial has to be utilized and as a result the range is drastically reduced to perhaps only a few miles. In both instances, the position of the lifeboat or dinghy could be such that the signals received by coastal stations would be so weak that accurate bearing information could not be obtained.

Survival under these conditions depends largely on nearby shipping or search aircraft and as a result the present trend, followed by the Royal Navy and Royal Air Force, is to rely more and more on the search and rescue method, utilizing S.A.R. Beacons working at v.h.f. and u.h.f., the significant feature of the method being the large search area which can be covered rapidly by an aircraft. The equipment available takes various forms but consists basically of a transmitter/receiver unit, battery and microphone/loudspeaker/aerial unit and because of its small overall dimensions and light weight, the complete equipment can either be carried in a specially designed belt to be worn by aircrew or stowed in a dinghy pack.

Use of the equipment by a survivor enables search craft to home in and effect rescue with the minimum delay. Range figures depend on the type of equipment used and the height of the aircraft. As a guide, an aircraft flying at an altitude of 10,000 ft can pick up a signal 60 nautical miles away representing an effective search area of about 11,000 square miles. At about 15 miles range, two-way speech communication can be established between the search craft and the survivor. By using more than one aircraft, sections of each search area can be overlapped so that the combined search effort can be concentrated in a specified area.

The equipment which operates in the v.h.f. and u.h.f. bands is usually supplied to operate on the International Distress Frequencies of 121.5 or 243.0 Mc/s and in use, the beacon transmits a distress signal which consists of a burst of carrier modulated by an audio tone. During the period between tone bursts, a reduced level of unmodulated carrier is transmitted. This facility, besides reducing the drain on the battery, also permits continuity of reception.

To use the equipment, the survivor releases a self-erecting aerial by pulling a cord attached to the aerial cover. The set is then automatically switched on in the beacon mode of operation and continues to transmit as a beacon for the duration of the life of the battery without any further effort on the part of the survivor. During beacon operation, the modulation tone is audible to the survivor. For R/T operation, two levers are provided on the transmitter/receiver unit, which can easily be operated by the survivor even in conditions of extreme cold with numbed and gloved hands. For R/T transmission, the user depresses a "transmit" lever and speaks into the microphone/loudspeaker unit. When a "receive" lever is depressed, the receiver background noise is audible and signals on the correct frequency and within range will be heard. When both levers are released, the set automatically reverts to beacon operation.

Erith, Kent. W. R. TIBBENHAM,
Burndept Limited.

"Transistor Audio Power Amplifier"

MAY I congratulate Messrs. Tobey and Dinsdale on their excellent article (November 1961 issue) describing a direct-coupled audio power amplifier. I am considering the use of such a circuit over the frequency band 40-80c/s in which one tends to lose a lot of power in the series capacitor C_7. I should like to have the author's views on the use of a direct coupling to the speaker by having a double rectified supply replacing, if you like, the batteries shown in Fig. 3. The crux of the problem is then the stability of the voltage level at the collector of Vt6. Would this be sufficient?

Cambridge. N. BETT.
(Electronics Section, Cavendish Laboratory.)

The authors reply:

The arrangement of the power amplifier described, with capacitive coupling to the load, was felt to be the most economical for music and speech applications.

The reduction in available output power due to the output capacitor is small at the frequencies mentioned by Mr. Bett, compared with the low-frequency reduction in output power in a conventional transformer-coupled amplifier. The voltage across the capacitor is in quadrature with that across the load at low frequencies, so that a value of (say) 1,000μF in series with a 15-ohm speaker gives at 40c/s:—

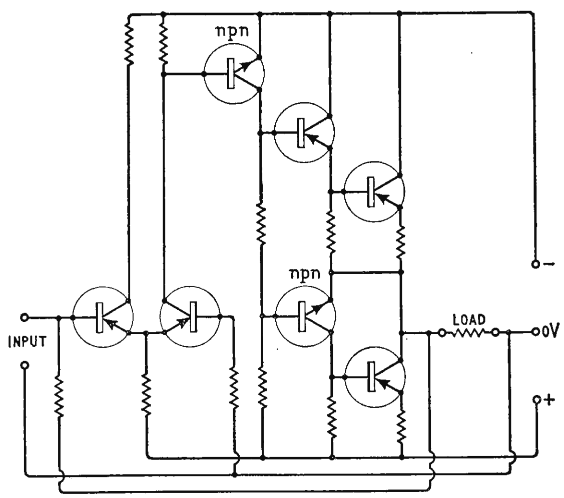
$$X_c = 4\Omega$$

Reduction in maximum available power is then:—

$$\frac{15^2 + 4^2}{15^2} - 1 \approx 0.3\text{dB } (8\%)$$

If the output capacitor is doubled it will cause a reduction of 0·075dB, which is less than 2%.

However, if operation down to zero frequency is re-



quired, and lower sensitivity, more noise and less economy in transistors can be tolerated, direct coupling to the load is possible with some rearrangement of the basic circuit as suggested in the figure.

It will be seen that a long-tailed input stage is used for the following reasons:—

1. To keep the offset d.c. through the load low under all conditions, including inequality of supplies and variations in ambient temperature (in the standard circuit the output-stage working point is slightly temperature-sensitive due to V_{be} variations in the single-transistor input stage).

2. In the standard circuit the reference point for the application of negative feedback is the positive side of the supply, so that if a centre-tapped supply is used, any ripple voltage between the earthy side of the load and the positive supply rail will appear across the load. In the circuit shown the earthy side of the load is the reference point for negative feedback so that the effect of any positive supply ripple is reduced by the negative feedback.

Incidentally, to get the lowest possible distortion in Version 2, it has been found desirable to put a resistor in series with the bootstrap capacitor C_6. A suitable value is 1kΩ.

With some transistors the high effective load impedance presented to Vt2 by the action of the bootstrap connection causes operation on a non-linear portion of the transistor characteristic at high voltages. The result may be second harmonic distortion which disappears when the bootstrap connection is disconnected.

The extra resistance in series with the bootstrap capacitor prevents the generation of second-harmonic distortion in this way, but the bootstrap connection still helps the amplifier to give the maximum negative swing without clipping of the waveform.

The effect does not occur on Version 1.

R. TOBEY.
J. DINSDALE.

Collectors Up or Positive Up?

NO doubt Mr. Baxandall (January issue, p. 23) is right, though his opponents are perhaps not so inexpert as he suggests. I feel, however, that his arguments are incomplete and invalid. My own demand for a *published* circuit is that it should be in a form which most nearly resembles all other published circuits of the same kind: it should be a passport photograph, a configuration by which it can be recognized quickly. Since we are unlikely to persuade the Americans and the Western Europeans, the Russians and for all I know the Chinese, to adopt the form he and, I suspect, you favour we need to accept two forms if his proposals are adopted. We also lose much of our stored knowledge of valve circuit configurations. This seems, in all, a high price for a small gain.

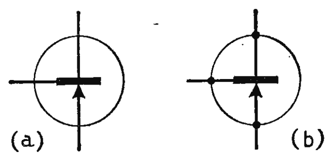
When we turn to consider circuits in detail, surely we redraw them if we must. In Mr. Baxandall's Fig. 10, for example, we should put the capacitor across the resistance to get it out of view and remove all evidence of earth from the circuit. We might even start with something like Fig. 1, page 17 of the same issue. Working drawings, however, can be redrawn in different ways to focus attention on different aspects, but they are drawn for oneself.

Communications are our business, communication is our problem. Surely, sir, we are the last people to insist on using our own private codes.

THOMAS RODDAM.

Transistor Symbols

ANY of your readers who may have followed my writings over the years should be in no doubt that I couldn't agree more with P. J. Baxandall in favouring the transistor symbol repeated here (a), and that in expressing his firm intention of continuing to use it he and I stand together as one man.



Many other writers in *W.W.*, plus M. G. Scroggie (Radio Laboratory Handbook, Principles of Semiconductors, and Foundations of Wireless, all with the same symbol) form a not negligible body of agreement, and I invite others to join. Admittedly the B.S.I. currently specifies the old point-contact symbol, but it is not impervious to enlightened change, as has been shown on several occasions.

Besides the reasons given by Mr. Baxandall for preferring (a), there is the advantage that in sketching circuit diagrams one can draw a recognizable version of it in a measured 1.3 sec., whereas the point-contact one takes longer. And (a) fits in better with a unified exposition of transistors and valves.

I suspect however that Mr. Baxandall and the other authors are not responsible for the version actually appearing in your January issue, (b). According to the universally accepted standard, it means that all the transistor electrodes are dead shorted together. The blobs are not only quite unnecessary (they would never be included in a sketched circuit) but quite incorrect.

As regards Mr. Baxandall's main theme—"earth down" or "negative down"?—I used to be torn apart by this dilemma, when all transistors were p-n-p and everyone knew a lot more about valves than transistors. But now, with both kinds of transistors available and students approaching transistors and valves together from scratch, I am convinced that Mr. Baxandall is right.

CATHODE RAY

[It has long been a *Wireless World* practice to use "blobs" to indicate connections to valve electrodes, and this has been extended to transistors. Rightly or wrongly we have also adhered to the practice of using "hop-overs" to leave no doubt when wires cross without a connection, so "Cathode Ray's" figure (a) could be wrong in *Wireless World* if he insists that the envelope is conducting.—Ed.]

"BELLING-LEE" NOTES

No. 37 *of a Series*

Some Mechanical Aspects of Design, Part 9.

To round off this subject, it is worth having a look at the means by which the execution of designs as finished products is assessed and regulated, i.e. inspection of manufactured goods. Inspection is necessary at all stages, from the incoming materials, through the piecepart and finishing processes, to final inspection of the assembled end products. To cope with the increasing complexity of modern electronic components, and the ever more exacting technological requirements of a vast range of products which now numbers nearly 2,000 different items, a highly diversified inspection organisation is necessary. The products range from simple terminals and connectors to highly sophisticated forms, and circuit protection components (fuses and cut-outs), interference suppressors, television aerials, and complete broadcast relay equipment. Inspection is carried out by a number of departmentalised units, each equipped for all the prescribed routines of examination appropriate to the different levels of manufacture, and much of the test gear required for carrying out the more specialised electro-mechanical test routines is produced in its own instrumentation section, which is also responsible for checking and maintenance of all instruments, both for performance and safety in handling.

We start with Goods Inwards Inspection, where raw materials are checked for dimensions (where applicable) and quality. In the case of sheets, rods, and bars, for example, this involves gauging the diameter, width, thickness, etc., and tests may be made of the hardness, and bending characteristics of metals, and so on. Mouldings incoming from other factories are checked for cleanness of finish, freedom from warping, dimensional accuracy, effectiveness of sealing where this applies, etc., and pieceparts arriving from other Works are similarly minutely inspected for accuracy and finish. Chemical analysis is carried out in the firm's Chemical Laboratory, and insulation and resistance tests are performed in the Electrical Testing Department.

Progressing to the factory, where the first stage involves turning the raw materials into millions of pieceparts, it would be impracticable to check every item fully to make sure that it complied with drawings, for this would require an army of inspectors which might possibly outnumber the production manpower, with precision gauging equipment, the cost of which could well exceed that of the manufacturing plant. Inspection is therefore carried out according to the latest scientific principles of Statistical Quality Control. The full theory behind this is beyond the scope of these notes, and readers who are interested in knowing more about it are referred to standard text books on the subject. Broadly speaking, however, it is based on controlled sampling of the products, whereby the chances of any fault escaping detection are reduced to a predetermined level, and, if the process is sufficiently selective, this represents a negligible proportion. In addition to revealing obvious faults, such as those caused by a broken tool for example, the significant feature of this method is that it shows up trends of aberration due to gradual wear before permissible tolerances are exceeded, and corrections can be made in time to prevent the occurrence of wasteful rejects. To be effective, Statistical Quality Control takes place in the Workshop where the job is being done, and is applied throughout all stages of manufacture, from making the pieceparts to putting them together in the sub-assemblies, and final assembly of the finished products. The Technical Inspection Department makes a further spot check on random samples taken from the finished batches before they are passed to Despatch Stores.

100% inspection of finished products is carried out on Government stores, and on items requiring certified release. It is also performed on products requiring adjustment after manufacture, such as the time setting of automatic cut-outs, or on components which do not lend themselves to mechanical gauging, such as the resistance of fuse-links, where variations of as little as four milliohms are examined, or on products where human safety is involved, such as the electrical isolation of set-top aerials.

Tests in the last named category are carried out in the Electrical Testing Department, which is equipped for all normal requirements such as the measurement of contact resistance and electrical breakdown, testing insulated wire for pin-holes, checking the characteristics of attenuators, diplexers, and R.F. amplifiers, with facilities for testing special components at up to 35,000 volts. And behind all this lie the scientific resources of the Type Testing Laboratories, some of whose functions have been mentioned earlier in these notes.

FREE

"Some mechanical aspects of design" will shortly be available in booklet form, including some of the notes which appeared earlier in this series. Please write if you would like a copy, or copies. There will be no charge, nor follow-ups.

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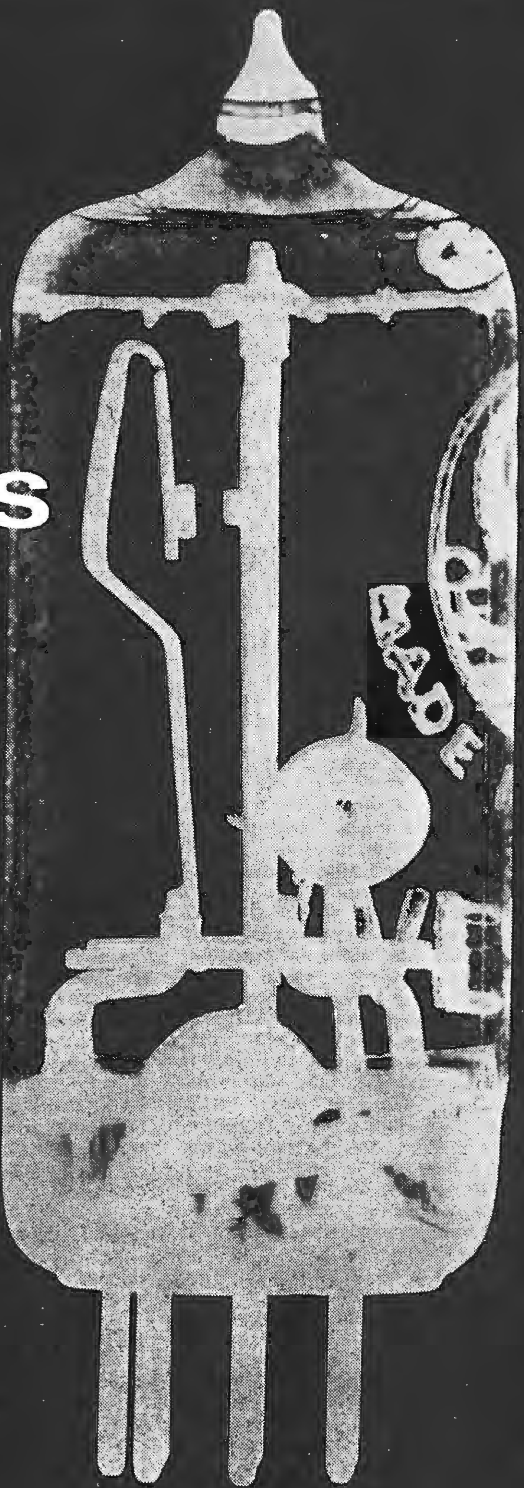

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Splitting the Load

OUTPUT STAGE USING TRANSISTOR WITH EMITTER AND COLLECTOR LOADS

By O. GREITER

THE natural approach of an engineer who is required to design an amplifier to deliver a fair amount of power is to begin by choosing a Class-B output stage. When the amplifier is to use transistors his move towards Class B is, if anything, even more pronounced. When he is called on to provide a very low output impedance his reaction is quite different. He may fly to positive feedback to help him out or he may adopt the cathode-follower or common-collector circuit but he will probably accept the inconvenience of Class-A working. In the more exacting cases this is not as bad as it seems, because frequently the requirement imposed on a transistor amplifier is that it should work satisfactorily up to a stated temperature which implies that the junction temperature will be very near its limit. Ingenious biasing circuits which use temperature-dependent biasing elements have been proposed for Class-B systems but these literally cannot be kept well enough in step with the junction to cope with practical operation even when, on steady-state tests, the tracking is good. The penalty is cross-over distortion, from which Class-A operation is the only escape.

The common-collector circuit would be ideal if it did not require so much drive: the base must be driven with a voltage which is roughly equal to the voltage across the load. Current is still required and it is found that the real problem has merely been moved back one stage. Even with valve stages it is often difficult to get sufficient swing at the grid to drive a cathode follower fully.

A common-emitter transistor can be driven very conveniently by a common-collector driver and this may in fact be achieved by forming the two into a compound pair. If this is done there is direct coupling between the two transistors: the leakage current of the first may upset the bias conditions of the second and, when this difficulty is avoided by a.c. coupling, we are faced by the problem of the very low impedance level.

The real trouble is that there is too much difference between the common-emitter connection and the common-collector connection. Each has its advantage, to excess, and each its disadvantage.

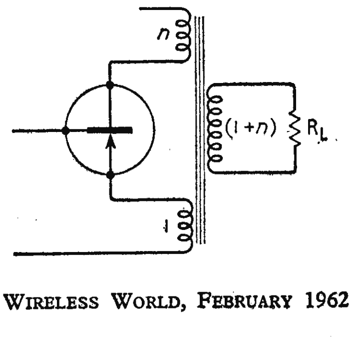


Fig. 1. Basic split-load stage.

We require a circuit in which we trade a loss in one characteristic against a gain in another. Fortunately such a circuit exists, although it seems to receive rather less than its due in the literature. This may be because it is possible to analyse it in a way which makes it almost incomprehensible: it can, however, be treated quite simply, so simply that one can only wonder why any other circuit is used.

Split Load Circuit

The skeleton of the circuit is shown in Fig. 1. It will be seen that instead of the primary of the output transformer being connected in the collector line of the common-emitter circuit, or in the emitter

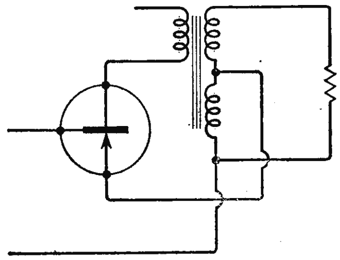


Fig. 2. Circuit of Fig. 1 re-drawn to show "application of feedback" principle.

line of the common-collector circuit, it is divided and the transistor is inserted at the division. By re-arranging the circuit in the form shown in Fig. 2 it becomes apparent that this is a way of providing negative voltage feedback round the output stage and it is well known that voltage feedback reduces the output impedance. It would appear to be a waste of time to go further, for the effect is known to involve a factor of $(1-\mu\beta)$. The difficulty is that μ is the unloaded gain and here there can be no doubt that even when the load resistor is removed the emitter path must constitute some sort of loading on the transistor. Referring, as one always does in these situations, to Bode* we see that we need to use the *fractionated gain* defined in Chapter V (5.11) and that this is the gain which would be obtained if it were possible to open the β circuit without affecting its impedance at either end. It may be that Mr. Roddam or Mr. Ray (Cathode variety) will care to show that this is easily done: it appears unlikely.

An approach which does not involve too much mathematics, and which will at least provide an approximation, begins by writing the a.c. equivalent circuit shown in Fig. 3a and then rearranging it

* *Network Analysis and Feedback Amplifier Design,* by Hendrik W. Bode, Ph.D., Van Nostrand Company Inc., New York, U.S.A. (1945).

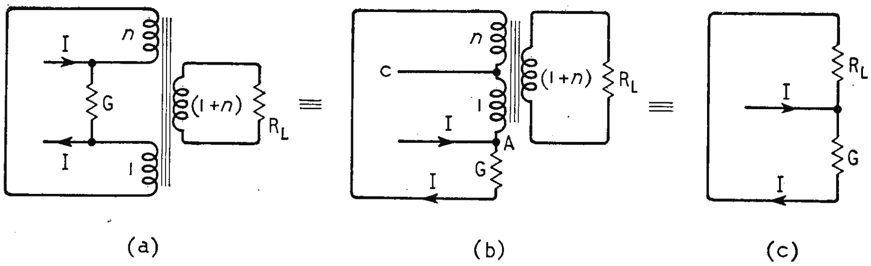


Fig. 3. (a) A.c. equivalent circuit of split-load stage (b) Rearrangement leading to (c) in which the term n has disappeared.

as in Fig. 3b. By taking the transformer secondary winding to have simply the ratio $(1+n)$ to the two windings n and 1 we get rid of n altogether in the final simplified form (Fig. 3c). Using the standard form for the transistor as a current generator we have the conductance G and we shall interchange between G and R $(=1/G)$, R_L and G_L $(=1/R_L)$ to suit the convenience of the analysis.

We see at once that the voltage which appears across the load is $I/(G+G_L)$. It will be convenient to work in terms of the transistor mutual conductance in producing a first approximation and we can write $I = g_m v_{be}$.

Across the section cA of the transformer we shall get a voltage of $I/(G + G_L)(1+n)$ and this must be added to v_{be} to find what the input voltage will be, thus

$$v_{in} = v_{be} + [I/(G+G_L)(1+n)]$$
$$= I[1/g_m + 1/(G+G_L)(1+n)]$$

This expression can be "processed" to obtain the gain but our first concern is with the output impedance. We remove R_L and from a source of infinite impedance we force in a current I_o, as shown in Fig. 4a. Turning this drawing round we find we can deal with the very simple form shown in Fig. 4b.

The voltage is $v = (I_o - I)R$ and the impedance must be v/I_o which we proceed to find. We have, as before, $I = g_m v/(1+n)$, for now G_L is zero and there is no external input supplied. Thus

$$v/I_o = R/[1+g_m R/(1+n)].$$

It is slightly more convenient to write this as an output conductance.

$$G_o = G + g_m/(1+n).$$

Typical Values

An indication of the order of magnitude may be useful at this stage. The Westinghouse WT10 series have an R of about 100Ω and a g_m of about 2A/V, or $1/g_m \approx 0.5\Omega$. If we take as a dividing line the point at which $G = g_m/(1+n)$ we see that as n falls below about 50 the second term becomes the dominating one. A typical load for such a transistor might be of the order of 10Ω and this could be matched by making $(1+n)/g_m = 10$ or, quite roughly, $n = 20$. By making $n = 4$ we should get an apparent source impedance of only 2.5Ω. We can thus move the impedance to any point we choose in the range 0.5 to 100Ω. The price we must pay is loss of gain. We can get an idea of the input impedance by writing $I = h_{fe} i_b$ and substituting this in a simplified form for the gain equation based on the fact that G is much smaller than G_L. Then

$$v_{in}/i_b = h_{fe}(1/g_m + 1/(n + 1)G_L.)$$

Generally we shall find that $1/(n + 1)G_L$ is sufficiently greater than $1/g_m$ for us to write $R_{in} = R_L h_{fe}/(n + 1)$ so that, as we might expect, the reduction of n leads to an increase of input impedance. When we have $R_L = 10\Omega$, $n = 4$ and $h_{fe} = 5$ we get an input resistance of 8Ω, for example.

A very simple approximation shows that the power gain is about $(1 + n)h_{fe}$, for the output voltage is multiplied by a factor of $(1 + n)$ if the mutual conductance is high enough while the current is multiplied by a factor of h_{fe}. This approximation is very useful in the preliminary stages of design, when the transistor types to be used must be selected. Either a limit can be set on n or the size of the driver transistor estimated as soon as the output transistor is specified. Attention can thus be concentrated on the essential of the design.

Approximations and Their Effects

A defect of the approximations made is that it has been assumed that the system is driven from a zero-impedance source. Although the transistor requires to have current fed into its base to provide the transistor action it is assumed that this current will be produced by the voltage indicated on the $v_b I_b$ characteristics. For the transistor we have selected the input resistance is only about 2Ω: only when the source resistance is well below this will the approximation be valid. An examination of the gain will show us what happens: the circuit to be considered is just the base-emitter path shown in Fig. 5, with an input signal v_{in}, a voltage at the emitter of $v/(1 + n)$, an external source resistance r_s, and a base current of i_b.

For this we have:—

$$i_b = [v_{in} - v/(1 + n)]/(R_s + R_{in})$$
$$I = i_b h_{fe} \text{ and } v = R_L I$$
$$= R_L h_{fe}[v_{in} - v/(1 + n)/(R_s + R_{in})] \text{ so that}$$
$$v[1 + R_L h_{fe}/(1 + n)(R_s + R_{in})] = v_{in} R_L h_{fe}/(R_s + R_{in})$$

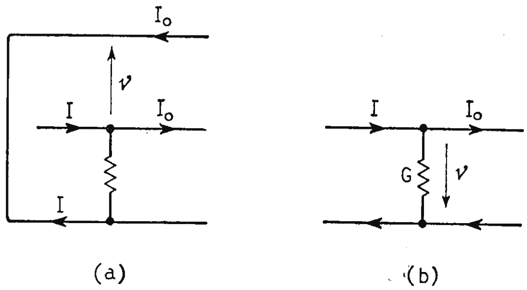


Fig. 4. (a) Current I_o is supplied from infinite impedance to determine output impedance. Form of circuit shown in (b) is simplified from (a).

Let us write $h_{fe}/(R_s + R_{in}) = g_m'$ and we have

$$v_{in}/v = 1/(1 + n) + 1/g_m' R_L \text{ or}$$

$$v_{in}/I = 1/g_m' + 1/(1 + n)G_L$$

When this is compared with the result obtained earlier we see that it differs only in the dropping of G, as it is small, and the substitution of g_m' for g_m. For practical purposes we may now assume that this same substitution must be made in the expansion for the impedance, without further proof.

One more factor merits examination. The feedback winding must deliver power and thus imposes a load on the transistor. This affects the value of G which should be used in the expression above and may make G large enough for it to be taken into account. We have the voltage $v/(1 + n)$ operating at a current level of i_b, so that we are concerned with $vi_b/(1 + n)$ which is $vI/(1 + n)h_{fe}$. The load power, however, is vI, so that unless $(1 + n)h_{fe}$ is small we need not concern ourselves too closely with this effect: if $(1 + n)h_{fe}$ is small we have other worries, for this is the approximate power gain.

The purist may by now be in rebellion at this patchwork approach. In small but exquisite writing he is preparing his polemic. Lay down your pen, Sir, touch not a hair of that old gray head†. None of these approximations is nearly so coarse as the approximations involved once we begin to put numbers into the expressions. Only accountants are happy if the books balance to the last penny when the company has lost thousands of pounds. The exact solution of this problem can be left to wait for the exact transistor.

† "The Ballad of Barbara Allen", which explains the spelling, and is quoted from a very shaky memory.

Effect of Winding Resistance

It is not without interest to note the effect of the resistance of the emitter winding on the output impedance. A very simple analysis shows that it adds a term $(g_m rG)$ to the output conductance, where r is the resistance in the emitter leg which is producing current feedback. In any practical case it is likely to be so small compared with G that this term will not lead us far from the very simple approximation $g_m/(1 + n)$. For example, a 1Ω resistance in the numerical case we have considered would represent a 20% correction when $n = 20$ and only 5% when $n = 4$.

The flexibility of this circuit, which makes it so attractive, makes it difficult to explore its possibilities adequately without considering a very wide range of examples. It will be clear that the freedom of choice of output impedance which the method offers provides the designer with a most valuable extra degree of freedom, and also leaves him with the painful task of deciding how best to use this freedom.

NEW YEAR HONOURS

AMONG the recipients of awards in the Queen's New Year Honours list are the following men in the world of wireless and electronics:—

Knighthood

Leon Bagrit, chairman and managing director of Elliott Brothers (London) Ltd., who is also chairman of the Electronics Board of B.E.A.M.A.

K.C.M.G.

Clive Loehnis, director, Government Communications HQ.

C.M.G.

J. B. Adams, lately director-general of the European Organization for Nuclear Research, who throughout the war was at T.R.E. working on the development of centimetric radar.

K.B.E.

A.V-M. Walter P. G. Pretty, Air Officer Commanding-in-Chief R.A.F. Signals Command.

C.B.E.

D. A. Barron, deputy engineer-in-chief at the Post Office.

Dr. E. G. Bowen, chief of the Division of Radiophysics in the Australian Commonwealth Scientific & Industrial Research Organization, who was in Watson-Watt's radar team.

J. O. H. Burrough, of the Government Communications HQ.

Dr. E. Eastwood, director of research, Marconi's W/T Company, and director of Marconi Instruments.

Capt. F. L. Millns, R.N., Captain of H.M.S. *Collingwood.*

O.B.E.

J. Bell, managing director, M-O Valve Company and vice-chairman of the British Radio Valve Manufacturers' Association.

F. J. Bentley, chief executive officer, Government Communications HQ.

Dr. R. A. Brockbank, staff engineer, Post Office Research Station.

Cdr. K. A. W. Pilgrim, R.N., British Joint Services Mission, Washington.

M.B.E.

H. D. Binyon, sales director, Solartron Electronic Group.

P. R. Keller, section leader, Marconi's W/T Company.

J. H. D. Ridley, head of B.B.C.'s Engineering Secretariat.

D. H. S. Simpson, chief telecommunications superintendent, G.P.O.

B.E.M.

R. A. Casey, engineering technical grade II, R.R.E., Malvern.

E. S. Ruff, radio supervisor, Government Communications HQ.

Dr. E. Eastwood (C.B.E.)

J. Bell (O.B.E.)

Automatic Tape Stop

By G. V. BUCKLEY,* A.M.I.E.E.

USE OF LIGHT REFLECTED FROM ADHESIVE METAL MARKER

It is often desirable for a magnetic tape or film machine to be stopped automatically at selected points of the programme and one effective method was described by B. H. Parks in the July, 1958, issue of *Wireless World* (page 308). The device to be described was developed for use with a magnetic tape recorder but could be used also on film machines.

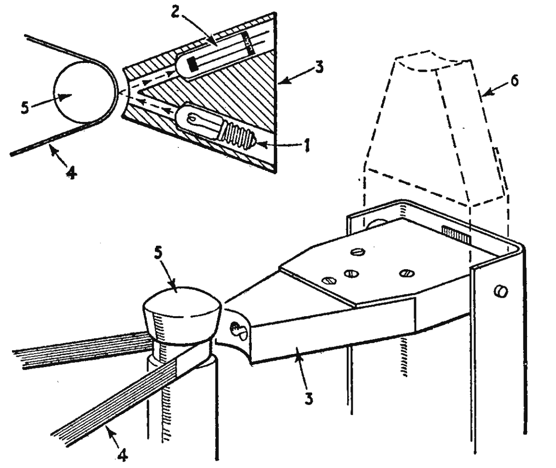


Fig. 1. Marker detector of automatic stop device.

A primary requirement for such a device in broadcasting is that the marker which is applied to the tape at the required stopping point should be easily applied and removed and should not damage the tape. This rules out methods employing holes in the tape or removal of the magnetic oxide from portions of the tape. Adhesive markers, however, are satisfactory from this point of view and have been used in the past to operate devices employing optical, capacitive or direct-contact methods.

Direct-contact methods employ metallic markers, either spliced into the tape or stuck on to it, which make contact with a stationary switch device, but, although this method is extremely simple, practical devices cannot easily be made really reliable under operating conditions. Optical methods have also been used in which the passage of a white or coloured adhesive marker on the tape is detected by diffuse reflection of light from its surface, but similar adhesive material is normally used for making joints in the tape so that it becomes difficult to make a detecting device distinguish between markers and joints, even if different colours are used for each.

These disadvantages and weaknesses can be overcome by using a metallic adhesive marker and making use either of its highly reflective properties for optical methods or its electrical properties for capacitive methods. The method finally adopted for this device uses an optical system, employing as a marker self-adhesive tape normally used for jointing video recording tape. The particular tape used for this development is made available to the B.B.C. by the Minnesota Mining & Manufacturing Co., Ltd., and is known as the "Scotch" brand of Aluminised Splicing Tape (Type No. 390/VR/¼in). (An alternative would be the same company's Aluminium Sensing Tape, Type No. 51-7/32S.) It is easily and quickly applied and removed, and does not harm the tape in any manner. There are other types of self-adhesive reflecting metallic tapes on the market which no doubt would be suitable for operating this

*British Broadcasting Corporation.

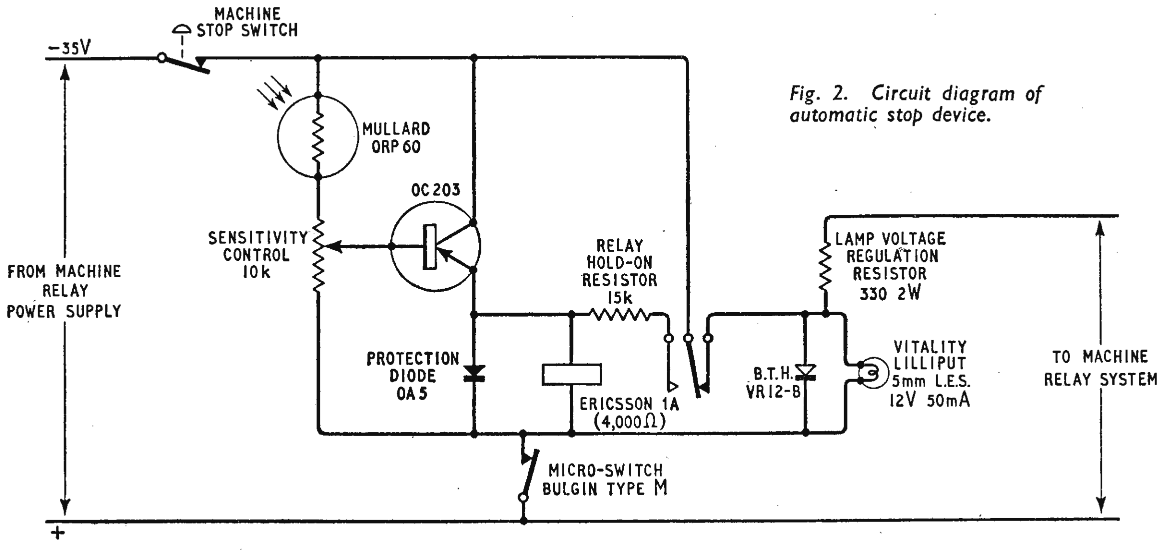


Fig. 2. Circuit diagram of automatic stop device.

device, but before using them it should be ascertained that the adhesive does not "creep" under pressure as, if it does, adjacent layers of the magnetic tape will stick together with unfortunate results.

The detecting device, together with a sectional view of the head block, is shown in Fig. 1. A light source (1) and a photoelectric cell (2) are housed in tunnels drilled into the solid head block (3). The two tunnels are drilled at angles such that the light from the lamp emerges from its tunnel and is reflected from the tape (4) back up the tunnel housing the cell. The geometry of the light path favours specular reflection and although the diffuse reflection from jointing tape will reflect some light on to the cell the ratio of light reflected from the metallic marker to that from jointing tape is great enough to ensure selective operation. Consistent results are obtained by keeping the distance between tape and head block constant and at its correct value, and this is achieved by choosing a point of reflection in the tape path where the position of the tape is fixed, such as at a pulley (5). The head can be swung to its off position (6) if it is not required or to facilitate tape threading. When the head is moved a short way from its operational position a micro-switch is actuated which disconnects the device entirely from the relay circuit of the machine. This feature also acts as protection against the device being operated accidentally by ambient light when being moved to its off position.

The device uses a cadmium sulphide type of photocell which has characteristics particularly suitable for this type of work. First, the photocell has a high value of sensitivity and a high current output enabling relay operation without intermediate amplification. Even where there are limitations of supply voltage and luminous intensity the cell can be used in extremely simple and reliable circuits for providing the necessary amplification, a typical example of which is shown in Fig. 2. Secondly, this cell has a response time (change of cell current time constant) of the order of tens of milliseconds, dependent on illumination, and this characteristic can be used for increasing the selectivity of the head by specifying the lengths of metallic markers and jointing strips to be used.

Very High Stage Gains

SELECTIVE AMPLIFIER STAGES USING HIGH-Q COILS

By G. W. SHORT

IT is possible, using a ferrite-cored inductor, to make a tuned circuit with a dynamic resistance of $50M\Omega$ at 1kc/s[1]. As this is much higher than the anode impedance (r_a) of any normal pentode valve, and as impedance transformation is possible in the tuned circuit, it is perhaps worth remembering that in these circumstances a stage gain in excess of the valve amplification factor (μ) can be obtained by matching the load to the valve. This has not been generally done since the days when low-μ triodes, used as audio amplifiers, provided much of the gain of a radio receiver and the meagre stage gains were augmented by using step-up transformers for inter-stage couplings.

A typical circuit is shown in Fig. 1, and to achieve maximum gain the transformer turns ratio (N) is chosen so that V1 sees a load equal to r_a. This is done by making $N=\sqrt{(R_L/r_a)}$ (In the present case R_L is the dynamic resistance r_d of the tuned circuit). The gain from grid to anode is $\mu/2$ when the anode load equals the anode impedance, so the overall gain becomes:

$$A_{max} = (\mu/2)\sqrt{(R_L/r_a)} \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

Putting in some typical quantities, suppose $\mu=6000$, $r_a = 1M\Omega$, and $R_L = 49M\Omega$. Then $A_{max}=21{,}000$, or 86dB.

The question arises: which is the better valve, one with high g_m and low μ, or one having high μ and low g_m? From equation (1),

$$(A_{max})^2 = \mu^2 R_L/4r_a$$

Writing μ/g_m for r_a:

$$(A_{max})^2 = \mu^2 R_L/(4\mu/g_m) = \mu g_m R_L/4$$

whence $A_{max} = \frac{1}{2}\sqrt{(\mu g_m R_L)}$

Evidently the best valve is one for which the product μg_m is high.

The table lists some practical values. (Valve manufacturers often quote only g_m and r_a for pentodes. Here it is easier to calculate $g_m^2 r_a$, which is equal to μg_m.) These suggest that, where voltage gain

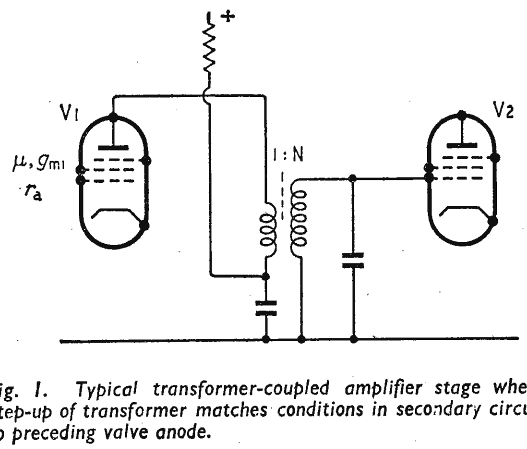


Fig. 1. Typical transformer-coupled amplifier stage where step-up of transformer matches conditions in secondary circuit to preceding valve anode.

Valve Type	g_m (mA/V)	r_a (MΩ)	μ	$g_m\mu$ (A/V)
6AC7	9	1	9000	81
6F1	9	0.9	8100	73
6J7	1.25	1.5	1900	2.3
6SH7	4.9	0.9	4400	22
6SJ7	1.65	1	1650	2.7
E180F	16.5	0.035	580	9.5
E810F	50	0.07	3500	175
EF50	6.5	1	6500	42
EF54	7.7	0.5	3800	30
EF86	1.8	2.5	4500	8
EF91	7.6	1	7600	58
EF95	5.1	0.69	3500	18
SP61	8.5	0.7	6000	50

alone is the criterion, the more old-fashioned types of television pentode are a good bet. In a practical amplifier however, noise and microphony may be the deciding factors and, if there is a very-high-impedance tuned circuit at both the anode and the grid, it may be necessary to select a valve with low grid-to-anode capacitance to avoid instability. However, an amplifier arranged for maximum gain by the technique described above is more stable than one with the tuned circuit connected directly in the anode circuit, without a transformer.

Gain v. Bandwidth

The price paid for high gain is reduced bandwidth, because there is a limit on the product of gain and bandwidth when a lumped-constant load is employed. This is of no consequence when the amplifier is required to be selective, provided that the tuning is not excessively sharp. (In the case of the coil mentioned in the reference, the working bandwidth (−3dB) would be about 13c/s at 1kc/s.) Applications which suggest themselves are a.c. bridge-detector amplifiers and selective amplifiers in harmonic-distortion-measuring instruments. Such equipment is nearly always mains-operated and physically large, so there is no particular advantage in using transistors. Indeed, the valve is at a distinct advantage, since it can produce a much greater stage gain. (In the case of a transistor *current* amplifier the maximum current gain is not realizable, because the following transistor presents a finite impedance. To get the maximum gain, its input impedance should be zero. In the case of a valve, where voltage amplification is required, it is possible, with care, to keep the input impedance of the following valve high, even compared with 50MΩ.)

If a high *stable* gain is required, the performance of the valve can be improved, without reducing the gain by stabilization of the anode current (which keeps g_m constant). The easiest way to do this is to use a cathode-bias resistor of higher-than-normal value and apply a stable positive potential to the grid to bring the anode-current back to its normal value (Fig. 2a). Then $I_k \approx v_b/R_k$.

Selectivity can then be improved by replacing the cathode bypass capacitor C_k by an acceptor circuit which presents a low impedance at the wanted frequency, or by incorporating a parallel-tuned rejector circuit (Fig. 2b). In either case, the component values should be chosen so that the impedance of the tuned circuit is $\ll(1/g_m)$ at the wanted frequency and $\gg(1/g_m)$ at the unwanted frequency.

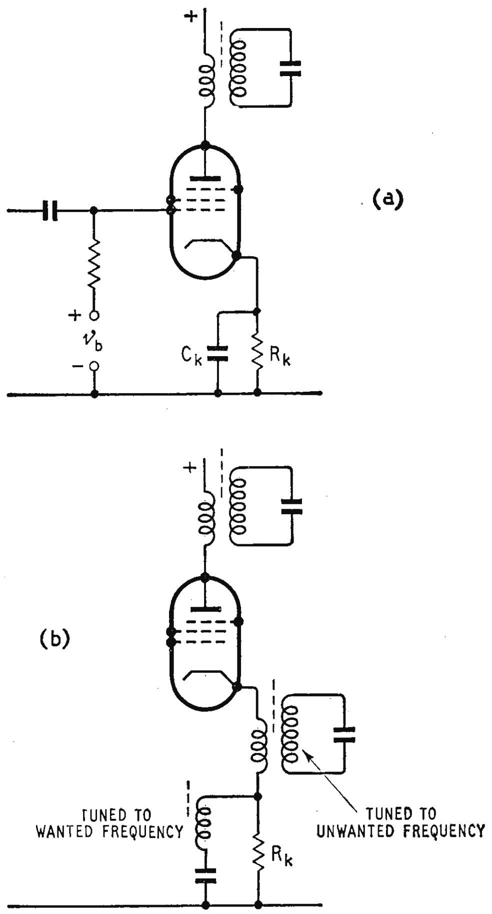


Fig. 2. (a) Use of high value cathode resistor for stabilization of anode current and g_m: v_{gk} is returned to normal by positive potential applied to grid. (b) Improvement of selectivity by use of two trap circuits in cathode lead.

The relevant mutual conductance here is that of the valve connected as a triode, which is usually about 25% greater than the pentode mutual conductance.

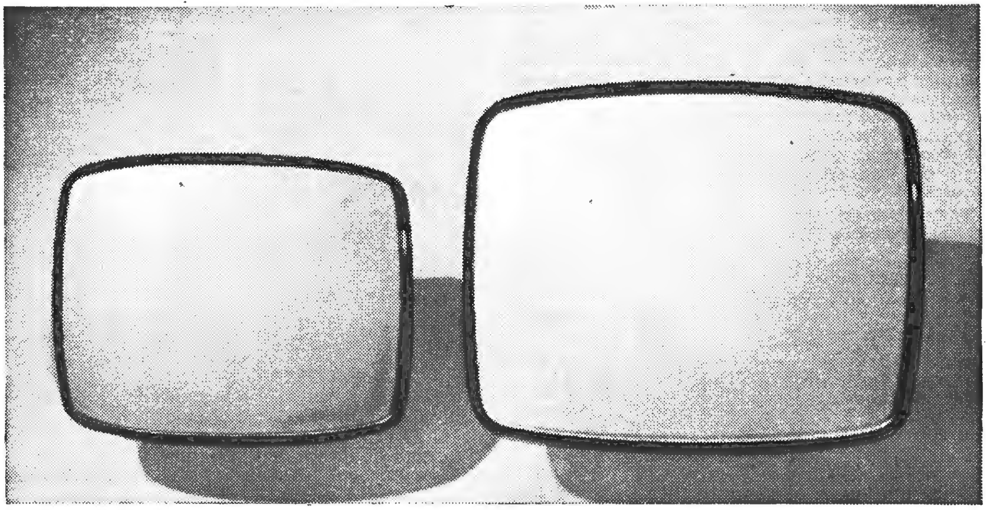
This trick should be useful in distortion measurements when a relatively weak harmonic signal has to be separated from a relatively strong fundamental, because it increases the signal-handling capacity of the valve at the unwanted frequency and so discourages the creation of spurious harmonics which might give misleading results.

REFERENCE

[1]J. M. Parkyn, " Extended Q-Meter Measurement and Measurement Accuracy." *Marconi Instrumentation*, September 1961, p. 61.

The coil taken as an example is one of inductance 50H, with a Q of 150, and it is described as " a practical component using ferrites". (Manufacturers' data for ferrite pot cores indicate that, using the largest cores, inductances running into hundreds of henrys can be made, using very fine wire. Despite the lower Q it would seem possible to achieve dynamic resistances of several megohms even at 100c/s.)

NEW 19" AND 23" TELEVISION CATHODE RAY TUBES



EDISWAN MAZDA TYPES
CME1901 AND CME2301

The CME1901 and CME2301 are, respectively, 19in. and 23in cathode ray tubes using magnetic deflection and electrostatic focus. The diagonal deflection angle of CME1901 is 114° and that of CME2301 is 110°. The shape of these tubes differs from the shape of conventional 110° tubes in that the face plates are more nearly rectangular. In addition the radii of curvature of the faces of the 19in. and 23in. tubes are greater than those of the 17in. and 21in. tubes. These changes result in a more pleasing presentation of the picture.

The external shape of the glass in the deflection region of these tubes is identical to that of conventional 110° tubes, enabling coils with conventional 110° internal mechanical contours to be used.

With equal values of final anode voltage, beam deflection in the CME1901 and CME2301 can be carried out with no more power consumption than in the CME1703 and CME2101.

GENERAL DETAILS

Rectangular face	Aluminised screen
Electrostatic focus	Silver activated phosphor
Magnetic deflection	Grey glass
Straight gun—non ion trap	External conductive coating
Heater for use in series chain	

Heater Current (amps)	I_h	0.3
Heater Voltage (volts)	V_h	12.6

TENTATIVE RATINGS AND DATA

Design Centre Ratings		**CME1901**	**CME2301**
Maximum Second and Fourth Anode Voltage (kV)	$V_{a2,a4(max)}$	17	17
Minimum Second and Fourth Anode Voltage (kV)	$V_{a2,a4(min)}$	14	15
Maximum Third Anode Voltage (volts)	$V_{a3(max)}$	±700	±700
Maximum First Anode Voltage (volts)	$V_{a1(max)}$	500	500
Maximum Heater to Cathode Voltage—Heater Negative d.c. (volts)	$V_{h-k(max)}$	180	180

Inter-electrode Capacitances (pF)		**CME1901**	**CME2301**
Cathode to All*	c_{k-all}	5	5
Grid to All*	c_{g-all}	8	8
Final Anode to External Conductive Coating (approx.)	$c_{a2,a4-M}$	1500	2000

*Inter-electrode capacitances including AEI "Clix" B8H holder VH68/81 (8 pin).

TYPICAL OPERATION

		CME1901	**CME2301**
Second and Fourth Anode Voltage (kV)	$V_{a2,a4}$	16	16–17
First Anode Voltage (volts)	V_{a1}	450	450
Third Anode Voltage for Focus-Mean (volts)	$V_{a3(av)}$	180	180
Grid Bias for cut-off of Raster (volts)		38 to 72	38 to 72
Average Peak to Peak Modulating Voltage for Modulation up to 350μA (volts)		34.5	34.5

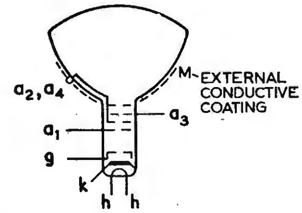
Note: All voltages given with respect to the cathode.

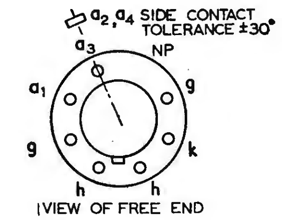
Maximum Dimensions (mm)	**CME1901**	**CME2301**
Overall Length	322	386
Face Diagonal	476†	598†
Face Width	420†	524†
Face Height	342†	422†
Neck Diameter	29.4	29.4

†The maximum dimension at the face seal may be 3.5 mm larger than this dimension but at any point around the seal the bulge will not protrude more than 2 mm.

Tube Weight (lb.)	**CME1901**	**CME2301**
Nett (approx)	13.5	27

Side Contact: CT8 (Cavity) **Base:** B8H



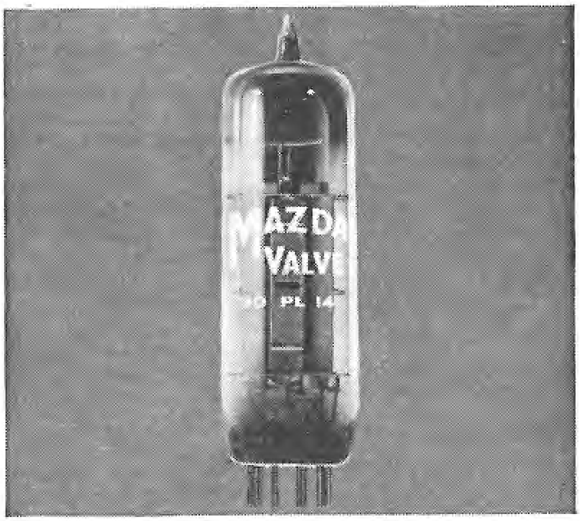


MAZDA VALVE COMMERCIAL DIVISION 155 Charing Cross Road, London WC2 Phone: GERrard 9797

EDISWAN EXPORT DIVISION Thorn House, Upper St Martin's Lane, WC2 Telex: 21521 Thorn London

THORN—AEI RADIO VALVES & TUBES LTD

NEW HIGH POWER TRIODE BEAM TETRODE



EDISWAN MAZDA 30PL14

The 30PL14 Triode Beam Tetrode is intended principally for use in frame deflection circuits of television receivers using 110° and 114° cathode ray tubes. The tetrode section has been designed to operate with a low ratio of screen to anode current whilst the available peak anode current and anode dissipation are higher than in the 30PL13. The higher available peak anode current will, in many cases, permit economies in output transformer design. The triode section is a general purpose triode with identical characteristics to the 6/30L2, and that of the 30PL13, for use as a deflection drive voltage oscillator, frame sync. pulse separator, etc.

Heater Current (amps)	I_h	0.3
Heater Voltage (volts)	V_h	16

TENTATIVE RATINGS AND DATA

Maximum Design Centre Ratings

		Triode	Tetrode
Anode Dissipation (watts)	$p_{a(max)}$	1	8
Screen Dissipation (watts)	$p_{g2(max)}$	—	2
Peak Anode Voltage (Pulse positive) (kV)		—	2*
Peak Anode Voltage (Pulse negative) (kV)		—	0.5*
Anode Voltage (volts)	$V_{a(max)}$	250	250
Screen Voltage (volts)	$V_{g2(max)}$	—	250
Heater to Cathode Voltage (volts rms)	$V_{h-k(max)}$ rms	150†	150†
Mean Cathode Current (mA)	$I_{k(max)}$	—	75
Resistance Grid 1 to Cathode			
Self Bias (MΩ)		—	2
Fixed Bias (MΩ)		—	1

*Maximum pulse duration 5% of one cycle with a maximum of 1 msec.

†Measured with respect to the higher potential heater pin.

TRIODE CHARACTERISTICS

Anode Voltage (volts)	V_a	200
Anode Current (mA)	I_a	10
Grid Voltage (volts)	V_{g1}	−7.7
Mutual Conductance (mA/V)	g_m	3.4
Amplification Factor	μ	18

TETRODE OPERATION IN FRAME TIME BASE

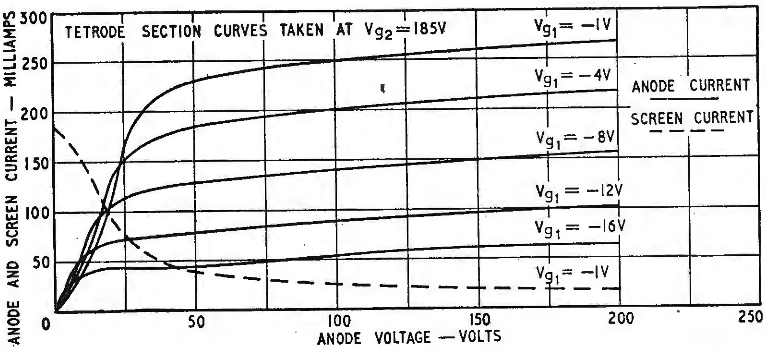
Allowance must be made in circuit design, not only for component variation, but for valve spread and deterioration during life. Values of total tetrode peak anode current, for an average valve when new and at the assumed end of life point for any valve, are as follows:—

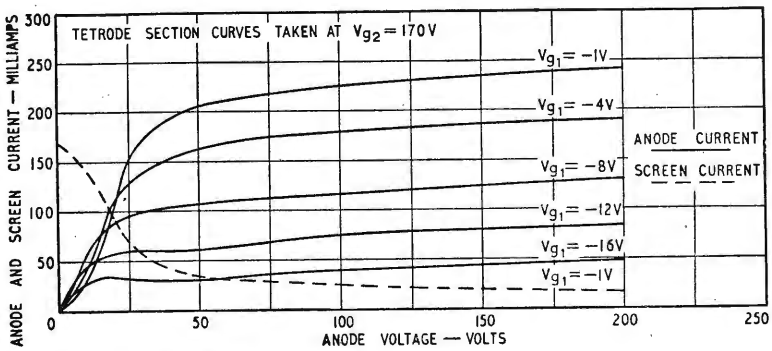
	V_a (V)	V_{g2} (V)	V_{g1} (V)	I_a (mA)
Average New Valve	55	170	−1	210
Assumed End of Life Condition	50	170	−1	135
Average New Valve	55	185	−1	235
Assumed End of Life Condition	50	185	−1	151

Mounting Position: Unrestricted

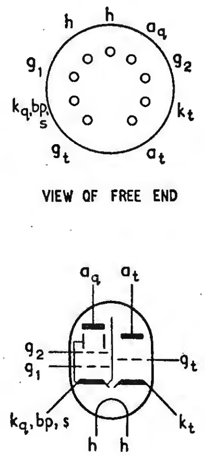
Base: B9A (Noval)

Tentative Characteristic Curves of Ediswan Mazda Valve Type 30PL14





Connections



Maximum Dimensions (mm)

Overall Length	78.5
Seated Height	71.5
Diameter	22.2

TRANSISTOR AMPLIFIER PAIRS

USE OF NEGATIVE FEEDBACK

By F. BUTLER, O.B.E., B.Sc., M.I.E.E., M.Brit.I.R.E.

THIS paper is intended to give the theory of a certain class of feedback amplifiers employing a pair of transistors in a phase-reversing circuit of a type more commonly designed for use with thermionic valves. Salient characteristics of the amplifiers will be described and some measured performance figures will be given and compared with calculated results. The paper will conclude with a description of some practical applications, all of which have been tested experimentally.

It is assumed that the reader is familiar in a general way with such properties of feedback amplifiers as can be inferred from the standard gain equation $m = A/(1 + AB)$. From this it is clear that the voltage gain m, with feedback, approaches more and more closely to the limiting value $1/B$ as the product AB is increased, A being the inherent gain of the amplifier without feedback and B the feedback coefficient. In practice the feedback network is usually constructed from stable, linear, passive elements and in the limiting case the performance of the amplifier becomes independent of A and is determined solely by the feedback circuit. It becomes linear and distortionless. By contrast, if the feedback network includes reactive or non-linear components then the complete amplifier becomes frequency-dependent or non-linear in its action and will in fact assume the characteristics of the feedback circuit. To obtain useful gain with feedback (m large), B must be small. Since AB must be large in comparison with unity it follows that A must be very large indeed if the full benefits of feedback are to be secured. There would appear to be no particular difficulty in constructing a multistage amplifier to give high voltage gain but unfortunately the term A in the gain equation is a complex quantity, since all practical amplifiers exhibit some phase shift between the output and input voltage. A phase displacement of 180 degrees turns negative into positive feedback and instability will result if the amplifier loop gain exceeds some critical value at frequencies corresponding to a phase reversal. Herein lies the basic difficulty of feedback amplifier design.

Fortunately the problem is much simplified if feedback is applied over two stages only. With reasonable care it is possible to hold down the phase shift per stage to less than 90 degrees. Moreover, two tandem stages can easily produce a voltage gain of 1000 or more and this is sufficient to allow most of the theoretical advantages of negative feedback to be realized in practice. In what follows, two-stage amplifiers only will be considered and remarks will be confined to those in which there is a deliberate reversal of the phase of the output with respect to the input voltage.

The feedback system will be that shown schematically in Fig. 1, in which the positive and negative signs show the instantaneous polarities of the various voltages to be considered. In this diagram R_1 and R_2 form the feedback network while r represents the input impedance of the amplifier. At low and moderate frequencies r would be the input grid resistance of a valve amplifier or the input impedance of a transistor amplifier in parallel with such biasing resistors as are required to set the correct operating conditions for this stage.

The arrangement in Fig. 1 is best analysed by means of the equivalent circuit shown in Fig. 2.

Let m = voltage gain of the feedback amplifier,
= E_o/E, or $E_o = mE$.
A = intrinsic gain of amplifier without feedback,
= E_o/e, or $E_o = Ae$.

Kirchhoff's equations for the two loops give:—

$$E = R_1 i_1 + e, \qquad (1)$$

$$mE = R_2 i_2 - e, \qquad (2)$$

$$e = mE/A = (i_1 - i_2)r, \qquad (3)$$

From (3):—

$$i_1 - i_2 = mE/Ar, \qquad (4)$$

From (1) and (3):—

$$i_1 = (E-e)/R_1 = \left(E - \frac{mE}{A}\right)\frac{1}{R_1}$$

$$= \frac{E}{R_1}\left(1 - \frac{m}{A}\right) \qquad (5)$$

From (2) and (3):—

$$i_2 = \frac{mE + e}{R_2} = \left(mE + \frac{mE}{A}\right)\frac{1}{R_2}$$

$$= \frac{mE}{R_2}\left(1 + \frac{1}{A}\right) \qquad (6)$$

From (4), (5) and (6):—

$$\frac{mE}{Ar} = \frac{E}{R_1}\left(1 - \frac{m}{A}\right) - \frac{mE}{R_2}\left(1 + \frac{1}{A}\right) \qquad (7)$$

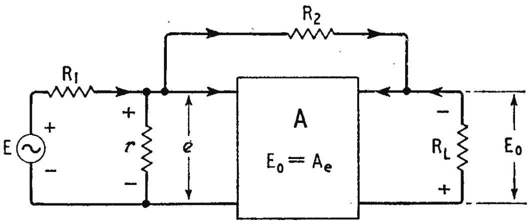


Fig. 1. Block diagram of feedback amplifier.

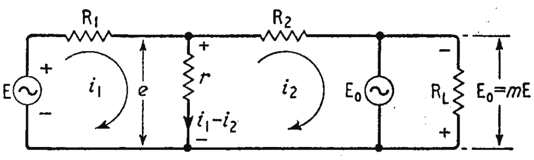


Fig. 2. Equivalent circuit of Fig. 1.

It is required to find an expression for m in terms of the other quantities. Equation (7) gives, after collecting terms and simplifying:—

$$E_o/E = m = \frac{AR_2/R_1}{A + 1 + R_2\left(\frac{1}{r} + \frac{1}{R_1}\right)} \quad .. \quad (8)$$

Note that E_o is opposite in phase to E so that, strictly speaking, the sign of m should be negative.

Some special cases can now be considered. Assume first that the amplifier gain A is infinitely large. The voltage gain is then $m = R_2/R_1$. This is simply the ratio of the feedback resistances and is independent of A. The actual values of R_1 and R_2 are decided by practical considerations. In the first place, R_2 should be large compared with the amplifier load resistance R_L in order to reduce its shunting effect on the load. For transistors with R_L of the order of 1 kΩ it is convenient to make R_2 about 10 kΩ. Once R_2 is fixed, the value of R_1 is settled by the desired value of m. If $m = 10$ and $R_2 = 10$ kΩ, then $R_1 = 1$ kΩ.

With an amplifier of finite gain A the voltage gain m with feedback is always less than the ratio

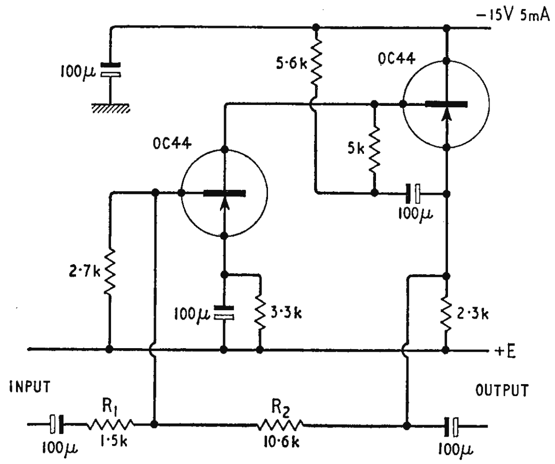


Fig. 3. Common emitter/common collector feedback amplifier.

R_2/R_1. Examination of equation (8) shows that the discrepancy will be small if A is very large in proportion to $1 + R_2\left(\frac{1}{r} + \frac{1}{R_1}\right)$. In practice, A will be of the order of 1000 and the inequality becomes $A \gg R_2\left(\frac{1}{r} + \frac{1}{R_1}\right)$. On the right-hand side the ratio R_2/R_1 is fixed by the required voltage gain m, leaving only the value of r to be considered. In a valve amplifier r will be about 1 megohm and its effect is negligible. For a common-emitter transistor amplifier r will be a few thousand ohms (comparable with R_2), so that R_2/r will not differ much from unity. Again, its effect is negligible. For a common-base amplifier, r may be quite small, say 25 ohms. If R_2 is 10,000 ohms the ratio R_2/r is 400 and this figure is quite large compared with the assumed value of 1000 for A. There will be a reduction of about 40 per cent in the observed value of m as compared with the "ideal" calculated value $m = R_2/R_1$. Finally, the effect of the ratio R_2/R_1 in the denominator of equation (8) must be taken into account. In practice this ratio will lie between 1 and 10 or so, depending on the gain required. If $A = 1000$ and $R_2/R_1 = 10$, the measured voltage gain will be about 99 per cent of the computed value, assuming for the moment that r is so large that its contribution can be neglected.

To sum up, a feedback amplifier of this type will have its properties determined largely by the two passive feedback elements R_1 and R_2 if:—

(a) The inherent gain A is very much larger than the ratio of the feedback resistances R_2/R_1.

(b) The amplifier input resistance r is so high that A is very large compared with R_2/r.

The input impedance seen by the signal source E is a little higher than R_1. If r is large the actual value is $R_1 + R_2/(A + 1)$. For some applications this relatively low input impedance is a disadvantage. One method of increasing it will be discussed later. It is not good enough merely to increase the value of R_1 since, to preserve constant gain, a proportionate increase must be made in R_2, leading to possible difficulties with shunt capacitance. Nor is it satisfactory to precede the feedback amplifier by a common-collector stage which will have a poor noise figure when operated from a high impedance source. It may also degrade the performance in other ways. Experience shows that the noise figure of most transistor amplifiers is a minimum when the input impedance is a few thousand ohms and when the amplifier is driven from a source having an internal impedance of the same order of magnitude. Fortunately, many signal sources can be designed to have this sort of output impedance.

Practical Amplifier Circuits.—Fig. 3 is the circuit diagram of a two-stage feedback amplifier on which some practical tests have been made. The component values shown are as actually measured, not just nominal figures. This does not imply that the choice is particularly critical. Tolerances of about five per cent are allowable if the absolute maximum undistorted output is not required. The nominal gain is set by the feedback resistances R_1 and R_2. The first amplifier is a common-emitter stage, unusual only in that the feedback resistance R_2 also forms one arm of a base-bias potential divider, the other arm of which is a 2.7 kΩ resistance. The output stage is a species of bootstrap circuit. Its thermionic-valve counterpart has been extensively discussed in *Wireless World*,[1, 2, 3, 4] and the transistor version calls for little further comment. Briefly, the action of the second transistor is such that the effective value of its base-bias resistance appears to be much higher than its actual physical resistance. As this component forms part of the collector load of the first transistor the gain of this stage is abnormally high. To achieve the same gain by standard R-C coupling would require impracticably high values of load resistance and supply voltage.

When constructing wideband feedback amplifiers it is good practice to use r.f. transistors with alpha cut-off frequencies of 10-20 Mc/s. If the response is to extend far into the video-frequency range, then v.h.f. transistors should be used. For the present

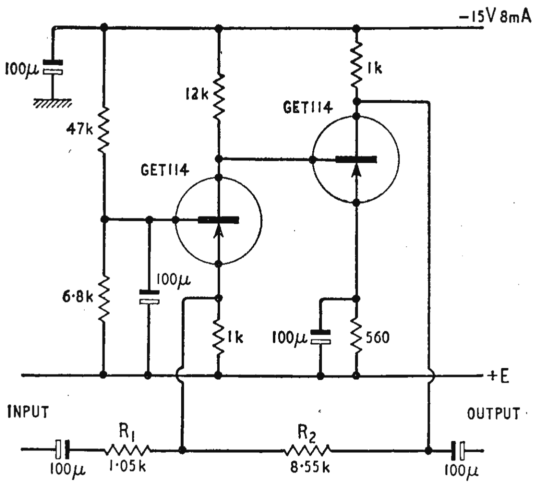


Fig. 4. Feedback amplifier with common base input stage.

application OC 44 transistors or similar types are entirely suitable.

Measurements on the amplifier gave the following results:—

Ratio of feedback resistances:	$R_2/R_1 = 7.07$
Overall voltage gain with feedback:	$m = 6.67$
Ratio:	$R_2/mR_1 = 1.07$
Inherent voltage gain:	$A = 1250$
Input impedance:	1.5 kΩ at 1000 c/s
Output impedance	14 Ω at 1000 c/s
Maximum open-circuit output voltage:	4V r.m.s.
Frequency response:	flat from 10 c/s to 22 kc/s
Value of R_1 to give unity gain, ($m = 1$):	9.9 kΩ

With a specially-filtered 1000 c/s signal source, adjusted to give 3V r.m.s. output from the amplifier, an attempt was made to measure the distortion products using a bridged-T LCR network to suppress the fundamental frequency. It proved impossible to measure the very low harmonic output because of instability of the null balance point, assumed to be due to random fluctuations of the test-oscillator frequency. Since the amplifier gain is reduced by feedback from 1250 to 6.67, i.e. by a factor of 187, it is reasonable to expect that the original distortion would be reduced by the same factor. Assuming 5 per cent distortion in the amplifier without feedback, this gives a figure of 0.025 per cent for the feedback amplifier.

Lack of suitable measuring equipment has made it impossible to give accurate gain figures at high radio frequencies but rough tests show that the response is well maintained up to at least 1 Mc/s. The power gain, working into a 1.5 kΩ load, is 16.5 dB, or 20 dB into 667 ohms.

Use of Common Base Input Stage.—A second amplifier has been constructed using the circuit shown in Fig. 4. The first stage is a common-base amplifier, characterized by a very low input impedance. The output stage makes use of a common-emitter amplifier to provide the requisite phase reversal. As before, R_1 and R_2 are the gain-setting feedback resistances. In essence the properties of this amplifier resemble those of the one just described, except that the output impedance is much greater. The measured gain, with feedback, corresponds rather less closely to the "ideal" figure determined by the ratio of the feedback resistances, in spite of the fact that the intrinsic gain is somewhat higher. This is due to the drastic shunting effect of the low input impedance of the first amplifier stage. From an examination of equation (8), this is only to be expected. If r is small, the numerical value of R_2/r is no longer negligible in comparison with A. Nevertheless the dominating effect of a large value of A is sufficient to permit the use of a common-base stage of very low input resistance.

A useful modification of the circuit is to replace the emitter bias resistance and by-pass capacitor of the second amplifier stage by a Zener diode of appropriate voltage rating. A 3.3-volt diode (Standard Telephones, Type Z2A 33), is suitable. It will serve to remove a troublesome low-frequency phase shift due to the usual CR bias network. Salient characteristics of the amplifier in Fig. 4, are:—

Ratio of feedback resistances:	$R_2/R_1 = 8.13$
Overall voltage gain with feedback:	$m = 6.82$
Ratio:	$R_2/mR_1 = 1.19$
Inherent voltage gain:	$A = 1500$
Input impedance:	1.05 kΩ at 1000 c/s
Output impedance:	163 Ω at 1000 c/s
Maximum open-circuit output voltage:	3V r.m.s.
Frequency response:	flat from 10 c/s to 22 kc/s
Value of R_1 to give unity gain:	7.05 kΩ

As before, there is useful gain up to 1 Mc/s.

An attempt was next made to modify this amplifier for operation down to zero frequency (see Fig. 5). For a start all capacitors were removed, including that between base and earth on the first transistor. The effect of this last operation is to introduce some negative feedback which reduces the internal gain of the amplifier. To offset this it

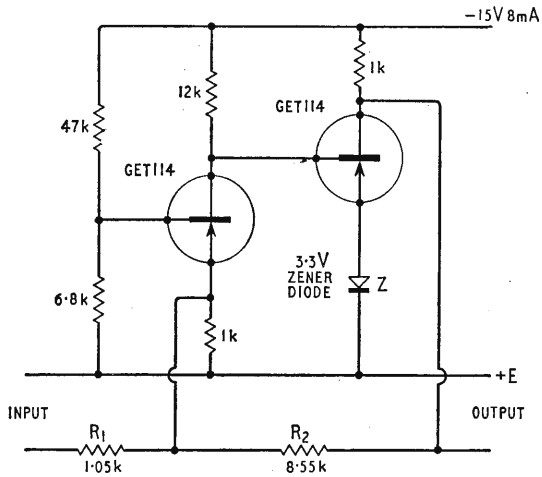


Fig. 5. Circuit of Fig. 4 modified for operation down to zero frequency.

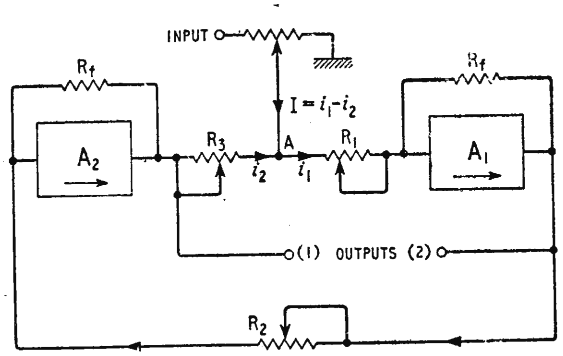


Fig. 6. Circuit with high or negative input impedance.

would be possible to use another Zener diode, assuming that one could be found with the correct operating voltage. Alternatively the resistance of the elements of the base-bias potential divider could be reduced, although this would result in drawing a much larger current from the power supply. The measurements quoted below were in fact made without changing the base-bias resistances.

One more precaution must be taken when using the circuit of Fig. 5 as a d.c. amplifier. This is to ensure that the input d.c. signal source and the output load are supplied with sufficient standing bias to maintain the proper quiescent conditions in the amplifier. Provided that the power supply is stabilized (say by using another Zener diode of 15 V rating), proper standing bias can be taken from the sliders of two potentiometers connected across the h.t. supply. Results of measurements on the circuit of Fig. 5 are as follows:—

Ratio of feedback resistances:	$R_2/R_1 = 8.13$
Overall voltage gain with feedback:	$m = 6.45$
Ratio:	$R_2/mR_1 = 1.26$
Inherent voltage gain:	$A = 345$
Input impedance:	1.05 kΩ at 1000 c/s
Output impedance:	170 Ω at 1000 c/s
Maximum open-circuit output voltage:	3V r.m.s.

The frequency response is exactly the same as that of the other two amplifiers.

Amplifiers With High Input Impedance.—For many purposes the low input impedance of the amplifiers described above is no disadvantage and indeed it can be turned to good account, particularly if the feedback network is proportioned to give an input resistance of 600 ohms. This figure is standard in telecommunications practice, filters, equalizers, attenuators and transformers being readily available to work at this impedance level.

On occasion, the low input impedance is inconvenient. It can always be raised to any desired figure by connecting an external resistance in series with the amplifier input but at the expense of reduced gain. A more elegant method has been described in a recent paper by E. Katell[5]. The principle is shown in Fig. 6. Two phase-reversing amplifiers are connected in cascade, the output of the second being fed back to the input of the first. The signal source is connected as shown. The variable resistances R_1 and R_2 serve to control the individual gains of the amplifiers A_1 and A_2, while R_3 permits adjustment of the output current of A_2. If i_1 is the input current of A_1 and i_2 the output current of A_2, the directions being as shown on the diagram, it is clear that the current I drawn from the slider of the input potentiometer is the difference between i_1 and i_2, or $I = i_1 - i_2$. If $i_1 = i_2$, no current is drawn from the source which thus sees an infinite output resistance, looking into the amplifier. If i_2 is larger than i_1 the input resistance is negative. This suggests that the system would be unstable but provided that the signal source has a sufficiently low resistance the net resistance between the amplifier input and earth remains positive and stability is assured. If R is the source resistance and $-r$ the negative input resistance the parallel combination is $-Rr/(R-r)$. This must always be positive to maintain stability so that, numerically, r must be greater than R.

Systems of this kind are notoriously difficult to set up and to be successful the amplifiers must be extraordinarily stable in gain and phase shift. They must also be virtually distortionless since, with reference to the signal source, i_1 is being partially or completely cancelled by i_2. If one of the currents contains distortion components not present in the other, cancellation of these will not take place. It is a tribute to the perfection of feedback amplifiers that the scheme can be made to work satisfactorily, even with a moderate degree of impedance multiplication. It is interesting to observe the amplifier output waveform as the point of infinite input impedance is approached. As the level of the input signal is increased the output waveform becomes progressively more and more distorted, particularly if the signal source is of high impedance. It is difficult to devise a more sensitive indication of amplifier linearity and with sufficient care in adjustment it might be possible to detect distortion as low as one thousandth of 1 per cent. This corresponds to the detection of distortion products which are 100 dB down on the fundamental level, a feat which is well-nigh impossible by any standard technique.

The negative-impedance amplifier can be used as an oscillator or Q-multiplier at frequencies between 50 c/s and 100 kc/s. For use as an oscillator the input gain control is removed and replaced by a suitable parallel-tuned LC circuit. The loop gain is then increased to the point where the negative input impedance is sufficient to neutralize the circuit losses. Sustained oscillation then occurs. Less than critical regeneration is used in the Q-multiplier and the signal to be amplified is also connected to the amplifier input, preferably through a high resistance. In either case (oscillator or Q-multiplier), outputs are available at points (1) and (2) in Fig. 6. The outputs are at low impedance and they are in antiphase. Their amplitudes are individually adjustable so that a balanced push-pull output is available if required.

An amplifier with infinite or negative input impedance is ideal for use after a diode demodulator stage. It is well known that distortionless demodulation of a deeply modulated a.m. signal calls for a near-unity ratio of d.c. to a.c. load resistances. This requirement is easy to meet and there is no trouble with instability because of the presence of the normal diode load resistance across the amplifier input. This scheme warrants a fuller investigation with reference to transistor circuits. Although uneconomical as regards components, use of the circuit would be justified where distortionless detection is required as in measuring equipment.

Miscellaneous Uses of Feedback Amplifiers.— The amplifiers shown in Figs. 3, 4 and 5 are ideal for mixing or combining audio signals from a variety of sources. Four programmes can be added, with negligible interaction, using the circuit of Fig. 7. The gain of each channel may be set or controlled independently of the others by suitable choice of the input resistance. The effective input resistance, looking into the amplifier from the junction point of R_1, R_2, R_3 and R_4 is accurately calculable. It is equivalent to a resistance $R/(A + 1)$ in parallel with the normal input impedance of the amplifier (e.g. the grid resistance of a valve amplifier or the input resistance of a transistor). If $R = 10\ k\Omega$ and $A = 1000$ the input resistance is close to 10 ohms, low enough to cause negligible interaction between the individual channel gain control settings.

A low-frequency crystal oscillator circuit is shown in Fig. 8. This is given more because of its scientific interest than for its practical utility since it is unnecessarily complicated. Two phase-reversing amplifiers A_1 and A_2 give an output which is in phase with the input. The quartz crystal forms the series element of a pi-network, the shunt elements being a low-value resistance and a thermistor, the purpose of which is to provide amplitude control. The feedback resistances are adjusted to provide enough loop gain to ensure reliable oscillation.

Finally, a brief reference may be made to the properties of amplifiers in which the feedback elements are reactive, non-linear or temperature-dependent. If either R_1 or R_2 is replaced by a pure reactance the amplifier behaves as an integrator or differentiator. A description of the basic principles of these circuits has been given in *Wireless World* by J. M. Peters[6]. Interested readers are referred to this paper.

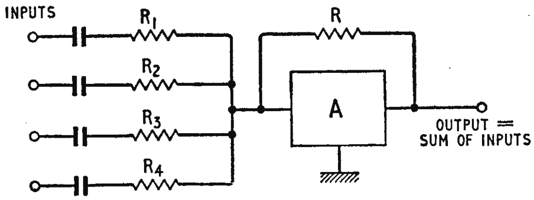


Fig. 7. Adding signals from four sources.

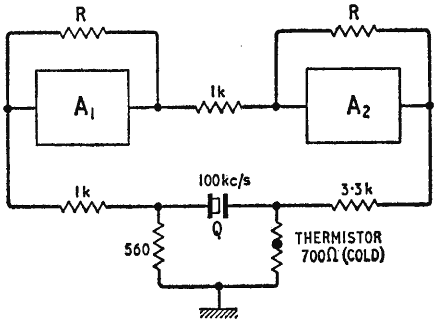


Fig. 8. Series resonant crystal oscillator.

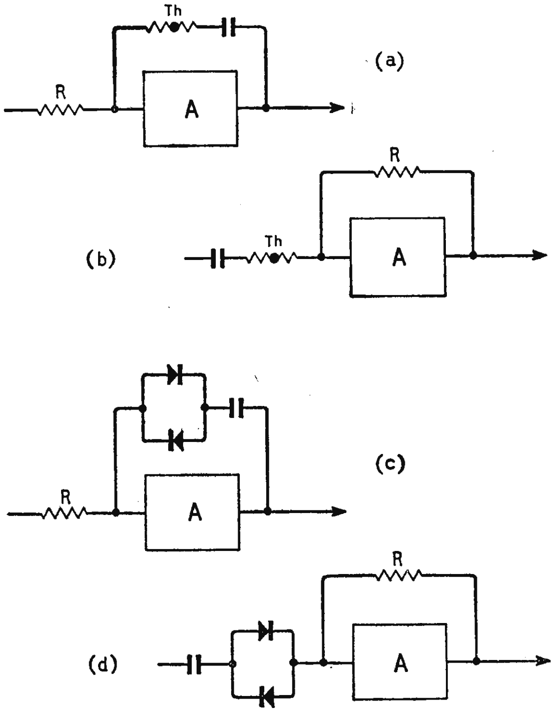


Fig. 9. Variable gain amplifiers.

In Fig. 9, schematic diagrams are given of four amplifiers which have applications in the field of control and regulation. Circuit (a) uses a thermistor as a feedback element. A rise in the input signal level causes an increase of energy dissipation in the thermistor. Its resistance falls and this causes a reduction of amplifier gain. The circuit can be used to control the output amplitude of an oscillator without causing distortion due to overloading and waveform clipping. Circuit (b) gives an increasing gain as the level of the input signal is increased. Neither circuit gives an instantaneous response to sudden level changes because of thermal lag in the control element.

Circuits (c) and (d) provide facilities similar to (a) and (b) respectively, but the response is instantaneous. They exploit the non-linear resistance of semiconductor diodes at low signal levels. Low-power silicon rectifiers are ideal for this application. Unfortunately both these circuits introduce some non-linearity distortion and they should not be used in high-quality audio equipment or in oscillators which are required to give a strictly sinusoidal output.

When thermistors or diodes are used as control elements it is essential to remove any standing d.c. by means of a blocking capacitor. These components are required to be responsive only to alternating signal currents. Thermistors should be protected from wide ambient temperature fluctuations since these may cause resistance changes large enough to swamp the changes due to electrical heating of the element.

In all cases where reactive or non-linear feedback elements are employed the control characteristics can be modified or "diluted" by the use of series or shunt resistance. In this way the overall

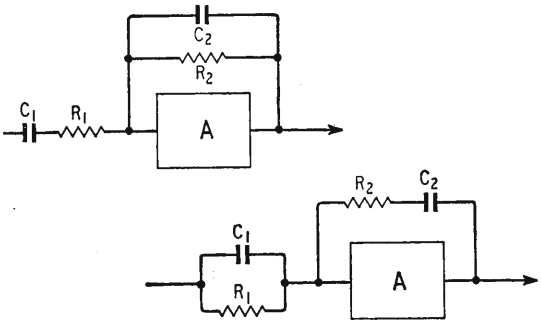


Fig. 10. Equalizer circuits.

gain characteristic can be tailored to suit almost any practical requirement.

Fig. 10 shows two amplifiers with associated equalizing networks. By proper choice of component values, any desired frequency characteristic can be provided. The list of examples given above is by no means exhaustive. Feedback amplifier pairs have been incorporated in RC oscillators, selective amplifiers, wave analysers, in measuring instruments and in pulse circuits. They are widely used in analogue computers, in servo mechanisms and in regulation or control equipment. In all these cases their small size and low power requirements are useful attributes but their chief attraction is that the performance comes close to the theoretical ideal.

Conclusion.—Transistor feedback amplifier pairs are easy to design when a.c. coupling is used between stages but are less simple when d.c. coupling is employed. Some care is necessary to establish simultaneously the correct operating points for both transistors. Minor errors in biasing will commonly result in one transistor being bottomed or cut-off and thereby rendered inoperative. The best way to set up the correct working conditions is to wire in permanently all the calculable components like load resistances, emitter resistances and feedback elements and to use variable components for the bias potential dividers. A signal input can then be applied and the output waveform examined on an oscilloscope. Adjustments can then be made to give maximum undistorted output, with symmetrical waveform clipping at the overload point. The makeshift components can then be removed, measured and replaced by fixed components of the nearest standard values. This procedure will not commend itself to the purist but it is a useful short cut. The alternative calls for a fairly precise knowledge of the transistor characteristics and it takes longer to measure these than to follow the line suggested.

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BOOKS RECEIVED

ARINC Transistor Specification Manual, published by ARINC Research Corporation, 1700 K Street, N.W., Washington 6, D.C., U.S.A., contains tables of transistor data and lists of transistors on ARINC and U.S. Military Preferred lists, reproduction of specifications including the American Mil-S-19500B and the British Joint Service Specification K1007 and commercial specifications. The manual is in looseleaf form so that it may be brought up to date or modified when necessary and contains 609 plus xi pages, price $10 per copy.

Guide to the Repair of Printed Board Assemblies. This booklet sets out in detail the procedures applicable to the repair of apparatus using printed-wiring boards. It should prove a most useful guide to the radio servicing apprentice and might offer a few tips to the established service engineer. Pp. 14; Figs 21. Electronic Engineering Association, 11 Green Street, London, W.1. Price 3s.

Second-class Radiotelephone License Handbook, by Edward M. Noll. Contains all the information required to pass the F.C.C. examination on operating and maintenance techniques. Over 650 questions are set with related text in additional chapters. American procedure is referred to throughout, but the technical information is, of course, applicable elsewhere. Pp. 240; illustrated. Howard W. Sams, and Co., Ltd., 2201, East 46th Street, Indianapolis 6, Indiana, U.S.A. Price $3.95.

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905:1959. *Interference Characteristics and Performance of Radio Receiving Equipment for Aural and Visual Reproduction.* Gives requirements and methods of testing to avoid interference in the long- and medium-wave bands, television and v.h.f./f.m. bands. Communal aerial systems are also mentioned. Pp. 50; Figs. 23. Price 12s 6d.

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FUNDAMENTALS OF FEEDBACK DESIGN

2.—STABILITY

By G. EDWIN

STABILITY is the very pith and marrow of all negative feedback problems. There are a number of different aspects of the stability question which are important to the practical engineer. It will be as well to note these before going on to discuss them in detail. First and foremost from the theoretical point of view there is the rule for determining whether a system will be stable. This is the foundation of all that follows but this is by no means sufficient for any constructor of hardware. Stable systems may be conditionally stable, in which case an overload may set them into an unstable state. The effect of component variations may be sufficient to modify the performance so that it becomes unstable, as many amplifier designers have found. Stability alone is not enough, for a system may be stable but still have a pronounced peak in its response which can cause ringing and overloading.

A proof of the rules which can be found is beyond our scope. The student who is sufficiently advanced will certainly have Bode's *Network Analysis and Feedback Amplifier Design* (Van Nostrand and Macmillan) at his elbow. Here it is probably appropriate to sketch out the way in which Wiener establishes a stability criterion in *Cybernetics* (M.I.T. Press and John Wiley). Suppose that we have a system which will give us, for an input $f(t)$ a delayed output $f(t - \tau)$, where τ is the delay. This is our amplifier path. At any time t we know the values of $f(t)$ for all lesser values of t: we cannot know what $f(t)$ will be for $t >$ now because someone may pull out the plug. The feedback path returns a portion of the output to the input. The first signal can be regarded as defined by $a_1f(t - \tau_1)$, while this first return will give us $a_2f(t - \tau_2)$. But this will in turn be recirculated. If you consider the signal to be a very short pulse which is simply delayed you will have no

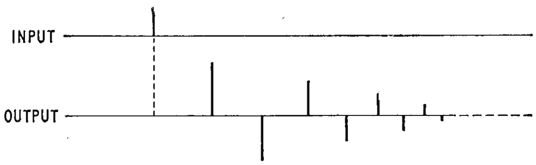


Fig. 2. Output of delaying feedback amplifier with very short single-pulse input.

difficulty in seeing that the result will be of the form shown in Fig. 2. Overall the result will be to give an output of

$$\sum_{k=1}^{n} a_k f(t - \tau_k).$$

The next stage in generalization leads to the more convenient form

$$\int_0^\infty a(\tau)f(t - \tau)d\tau.$$

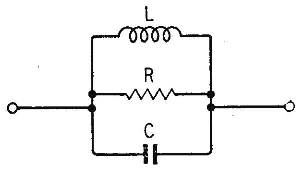
The important point about this expression is that the integration is from 0 to ∞ and not from $-\infty$ to ∞. Any physical system which converts an input $f(t)$ into

$$\int_{-0}^{\infty} a(\tau)f(t - \tau)d\tau$$

with $a(\tau)$ not zero for all negative values of τ is apparently predicting the future. Which of us would dare to do that?

A passive system which illustrates the problem is the simple low-pass filter. If we wish to provide band-limitation for a television signal we must

Fig. 3. Simple circuit for introducing idea of frequency functions in the complex plane.



produce from a pulse input an output of the form $(\sin x)/x$. Equalizer systems using delay networks can be made to give as good an approximation as you like to this, provided you are prepared to wait. The ideal system will have an infinite delay.

What actually happens when a system claims to tell the future is that it ignores the past. It may oscillate or it may lock over against a non-linearity: both of these effects are useful in their own way. An amplifier, however, must not set up in business for itself.

The analysis given by Wiener proceeds to convert the time function above into a frequency function. This can be plotted to describe a boundary dividing the whole plane into two regions which can be called inside and outside. In Wiener's treatment the point at infinity cannot be inside; for practical work we transform the function to get a more convenient rule.

This question of frequency functions in the complex plane seems to produce a psychological block in some engineers, possibly because it is followed usually by some steep mathematics. Consider, however, the very simple circuit shown in Fig. 3. This will frighten no one. The impedance of this circuit is Z, where

$$Z^{-1} = R^{-1} + (1 - \omega^2LC)/j\omega L.$$

If we now write $\omega_0{}^2LC = 1$ we get

$$Z^{-1} = R^{-1} + (1 - \omega^2/\omega_0{}^2)/(\omega/\omega_0)j\omega_0L$$

and we can then substitute

$$\Omega = (1 - \omega^2/\omega_0{}^2)/(\omega/\omega_0)$$

to get

$$Z^{-1} = R^{-1} - j\,\Omega/\omega_0L$$

so that

$$Z = R/(1 - j\,\Omega R/\omega_0L).$$

It is not too difficult to show that if we write $Z = x + jy$ and plot the values of x and y for variations in Ω we get the circle shown in Fig. 4. This circle is traced out on the end of the vector Z. At $\Omega = 0$, where $\omega = \omega_0$, Z is just R. At $\Omega = \infty$ where $\omega = 0$, Z is zero, as is also the case for $\Omega = -\infty$, where $\omega = \infty$. If $\omega < \omega_0$, y is positive,

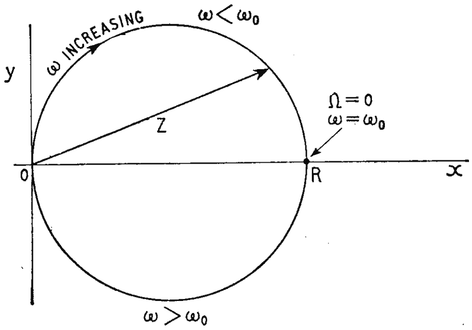


Fig. 4. Frequency plot in complex plane for varying ω and Ω of vector impedance Z of circuit of Fig. 3.

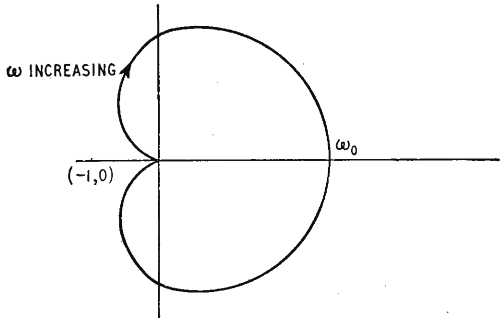


Fig. 5. Typical Nyquist diagram drawn in terms of $-\mu\beta$, the return ratio T.

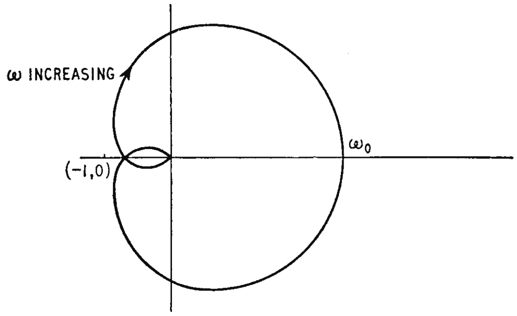


Fig. 6. Nyquist plot of amplifier which can become unstable if gain, μ, or feedback, β, increases too much.

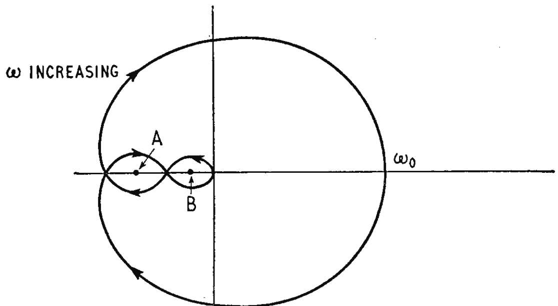


Fig. 7. Nyquist plot of conditionally stable amplifier. Is A inside, or B, or both or neither ?

so the boundary is traced in the direction of the arrow.

Fig. 4 is a frequency plot in the complex plane. The fact that we have used Ω instead of ω means little, because for each numerical value of ω we can find the corresponding Ω. There are even tables of $x/(1 - x^2)$ and $(1 - x^2)/x$ published. It is much easier to work with Ω , because then we keep special cases off the graph. We can see at once that the phase angle of Z will be $\pm$ 45° if $\Omega R/\omega_0 L = 1$, for example, and that the impedance is $R/\sqrt{2}$ at these two points. Since in ordinary language $Q = R/\omega_0 L$ we have the very simple expression $\Omega Q = 1$ for the bandwidth equation.

This figure gives a quite clear idea of what we mean by inside and outside. We must now apply this to the feedback amplifier. The mathematics is concerned with the position of the point at infinity. In the amplifier equation we have the form

$$\mu_f = \mu/(1 - \mu\beta).$$

By working with $(1 - \mu\beta)$ and assuming that all will be well when $\mu = 0$ (which is why we do not work with) $(1 - \mu\beta)/\mu$), we can now demand that the expression $(1 - \mu\beta)$ should not have the origin (0, 0) inside it. After all, if we turn everything upside down, the point at infinity must come down to (0, 0).

All this has been desperately simplified in an attempt to show how the full formal treatment proceeds and like all rough sketches it is neither rigorous nor strictly correct. The crux of the matter is, however, that if we draw a graph of $(1 - \mu\beta)$ in the complex plane, remembering that μ and possibly β are functions of frequency, the system will not be stable if the origin is inside the graph.

We need not plot $(1-\mu\beta)$ of course. We can plot $\mu\beta$ or $-\mu\beta$. Bode calls $(1-\mu\beta)$ the return difference or feedback, F; $-\mu\beta$ is the return ratio, T, and $\mu\beta$ is the loop transmission. If F is used, the origin is the critical point: if T is used the critical point is $(-1, 0)$: if $\mu\beta$ is used the critical point is (1, 0). As Bode points out, each valve produces a sign reversal and we usually look after the minus sign by using an odd number of valves or a suitable connection. After all, we know our target is negative feedback. Thus μ is normally negative and β positive and we can most easily work with T and let the mid-band signs look after themselves. Strictly speaking the Nyquist diagram is the plot of F which we have discussed: in practice we talk of the Nyquist diagram when referring to any of these three forms. A typical appearance for a Nyquist plot is shown in Fig. 5 and is drawn in terms of T. This seems to agree with the reasonable feeling that around ω_0, the middle of the working range, the quantity we are considering should be at its greatest. With this particular shape the system will always be stable.

Consider now Fig. 6. Remember that this is a plot of $\mu\beta$ (forget the minus sign) and that if we double μ or β the diagram will become twice the size. As the figure is drawn this will just make it enclose the point $(-1, 0)$ nicely, so that here we have a diagram of the kind we expect to find troublesome in practice: too much gain or too much feedback will lead to instability.

The diagrams shown in Figs. 5 and 6 are symmetrical about the real axis but this is by no means a practical necessity. In any normal a.c. coupled amplifier the characteristic will usually be considered independently in the two regions, below and above ω_0 , and quite different circuit elements determine the shape of the Nyquist diagram. A direct-coupled amplifier represents a rather different situation. In an a.c. amplifier the gain at zero frequency is zero and parasitic capacitances ensure that the gain at infinite frequency is also zero. For a d.c. amplifier we should get, for example, only the lower half of the apple of Fig. 5, with $\omega_0 = 0$. As $j(-\omega) = -j\omega$ we can show that by considering negative frequencies we shall generate the other half of the boundary, a mirror image of the half-diagram for positive frequencies. This will close the diagram up and solve the problem of what is inside and what is outside.

If you believe you can tell inside from outside, examine Fig. 7. Is A inside, or B, or both or neither? This is the type of characteristic which is known as conditionally stable. The test applied is to draw the vector joining the point under consideration, which will, in fact, be $(-1, 0)$, to the boundary and then to find through what angle this vector rotates as the boundary is traced. The result is shown in Fig. 8, from which it can be seen that going round the contour rotates the A vector through an angle of 4π, while the B vector gets back to where it started. Thus B is not inside the boundary at all. If the scale of the diagram is such that B is the point $(-1, 0)$ the system will be stable. In this arrangement, however, changing the scale by a reduction of gain will bring $(-1, 0)$ inside the boundary. The gain may be reduced by overloading and the system can then begin to oscillate and keep itself overloaded: the gain will be low when the supplies are first connected and again oscillation may begin. Although these conditionally stable designs have been used, they are suitable only for very carefully judged situations in which these factors can be taken into account.

While an understanding of the Nyquist diagram is necessary as a background, it is not a very convenient graphical form for design work. The polar plot records the complex value of $\mu\beta$ at all frequencies. If we want to change a single circuit element we must re-draw the whole diagram. This makes it rather hard to estimate what the effect of a change is likely to be. Let us suppose that we have two quantities $|A_1|\angle\theta_1$ and $|A_2|\angle\theta_2$. If we multiply these together the result will be $|A_1||A_2|\angle\theta_1+\theta_2$ An even easier form is $\log(|A_1||A_2|) = \log|A_1| + \log|A_2|$ and $\theta_1 + \theta_2 = \theta_1 + \theta_2$. We can therefore proceed by treating the gain in decibels (the engineer's way of using his logs!) and the angle of each circuit section independently as functions of frequency and then adding all these gain and angle components. This will provide us with results in the form in which we usually measure them and will, moreover, enable us to use very simple graphical methods for our design calculations. The fact that we now have two diagrams instead of one is not a disadvantage, but the fact that one of the diagrams is the familiar amplitude-frequency response makes it extremely easy to coordinate the design calculations with the experimental results.

Before we are through we shall have seen all the diagrams of this kind we could possibly want and both energy and space can be conserved by not including typical diagrams here. If we regard the Nyquist diagram of Fig. 6 as typical of the sort of behaviour with which we shall be dealing, we can see that there are two basic criteria which can be

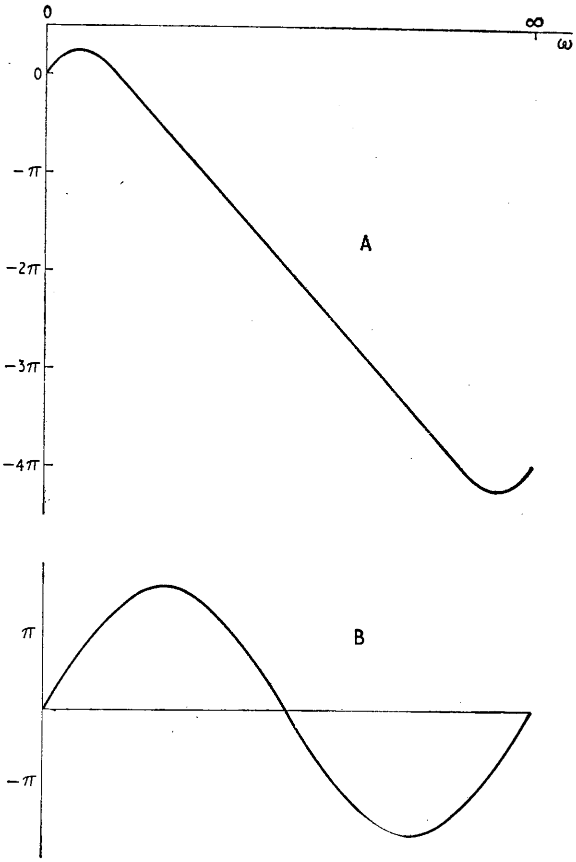


Fig. 8. Plots against ω of angles of vectors joining points A and B to boundary of curve of Fig. 7.

applied to each end of the characteristic. At the frequency at which the phase shift differs by 180° from the mid-band reference phase, the quantity $|\mu\beta|$ must be less than unity, so that $20 \log|\mu\beta|$dB must be negative. This quantity, without the minus sign, is the gain margin. At the frequency at which the value of $|\mu\beta|$ is equal to unity, so that $20 \log |\mu\beta| = 0$, the phase must not have come round through 180°. The amount by which it is short is called the phase margin.

Another very important factor is measured by the shortest distance on the Nyquist diagram from $(-1, 0)$ to the boundary. This is called the stability margin and is easy to calculate and even easier to measure. It is also of special interest in connection with the ringing on transients sometimes produced in feedback amplifiers.

At this stage we can regard the foundations as established. Although the discussion is elementary it is sufficient for the design of systems which are well beyond the range of the guess-and-see school. We shall now proceed to erect, brick by brick, a simple and practical method of, as it were, building the amplifier on paper.

Part 1: We regret that in some copies the first part of the expression on line 25 of p. 28 was not printed clearly: it should read $[-\mu\beta/(1-\mu\beta)]$. . . . (Ed.)

SUNSPOT CYCLES

TO WHAT EXTENT IS LONG-TERM FORECASTING RELIABLE?

By T. W. BENNINGTON

SUNSPOT cycles are of interest to radio engineers and amateurs concerned with long-distance communication because they are indicative of the variations in the activity of the sun and hence in those of the general level of ionization in the ionospheric layers, which are produced by the solar radiations. As there may be some readers who are new to this subject we may, perhaps, be permitted to elaborate a little.

The solar radiations alluded to are those emitted by the sun in the ultra-violet and X-ray part of the spectrum, and which, on entering the earth's atmosphere, expend their energy in ionizing the different gases which exist at ionospheric heights. These emissions are not constant, but vary according to the sun's state of activity, and one of the indications of this activity is the number of sunspots which appear upon the solar disc. This varies from day to day and from month to month, but it exhibits a marked long-period variation, such that the number and frequency of occurrence of sunspots rises from a minimum to a maximum and falls again to a minimum within a period of about eleven years, which variation constitutes a sunspot cycle. Thus the intensity of the solar radiations and hence the general level of ionization in the ionospheric layers also vary in sympathy with the sunspot cycle: at sunspot minimum the ionization is at its lowest and at sunspot maximum at its highest. The frequencies which are of use for long-distance radio communication are dependent upon the prevailing ionization: when this is high the higher frequencies are best, but when it is low they become unusable and lower frequencies have to be employed.

Some Characteristics of Sunspots

The average duration of the eighteen complete sunspot cycles for which records are available is 11.1 years: that is the average period which has elapsed whilst sunspot activity has varied from a minimum value, through a maximum and back to a minimum again. Thus the phenomenon is generally referred to as the eleven-year sunspot cycle.

The first spots of a sunspot cycle appear in solar latitudes of approximately 30° N or S. As the cycle continues there is a slow but definite change in the latitude of the spots and by the time of sunspot maximum the average latitude is about 16° N or S. This drift of the spots towards the solar equator continues throughout the cycle and towards the end of the cycle the spots belonging to it appear only a few degrees N or S of the equator. Thus the mean latitude in which the spots appear at any epoch is itself an indication of the progress of the cycle. Towards the end of the cycle spots again begin to appear in about latitude 30° N or S, and these are the first spots of the new cycle. The cycles in fact overlap each other by some two to three years, for the spots continue appearing in the four solar areas—two in high latitudes N or S of the equator, and two just N or S of the equator—for about that time. At the time of minimum they are about equally numerous in the four areas, after which the low-latitude spots gradually die away, and the old cycle finishes. These facts are of some assistance in predicting the time of sunspot minimum, and the definite start of a new cycle.

Another feature of interest concerns the magnetic polarity of sunspot groups. Sunspots most often appear in groups, consisting predominantly of two spots, which move with the sun's rotation from east to west across the photosphere (the visible solar surface). The most westerly of the two spots is called the "leader" and that in the easterly position the "follower." The two spots have strong magnetic fields which are of opposite polarity, as if an enormous horseshoe magnet were embedded in the sun with its poles protruding at the position of the spots.

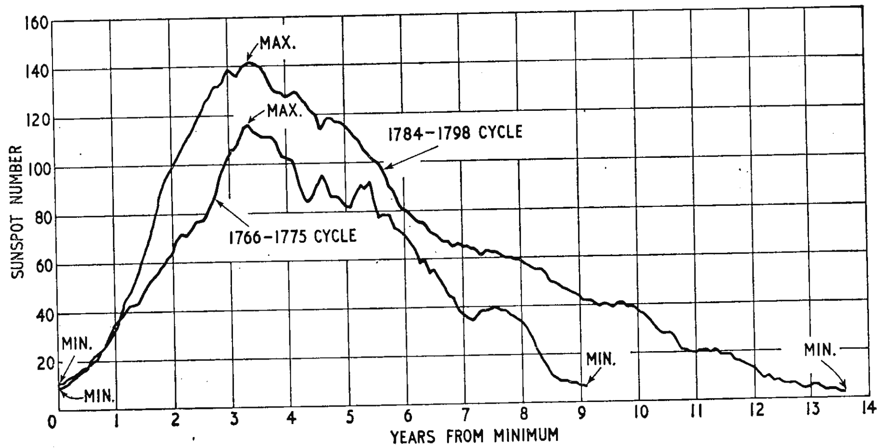


Fig. 1. The longest and the shortest sunspot cycles so far recorded.

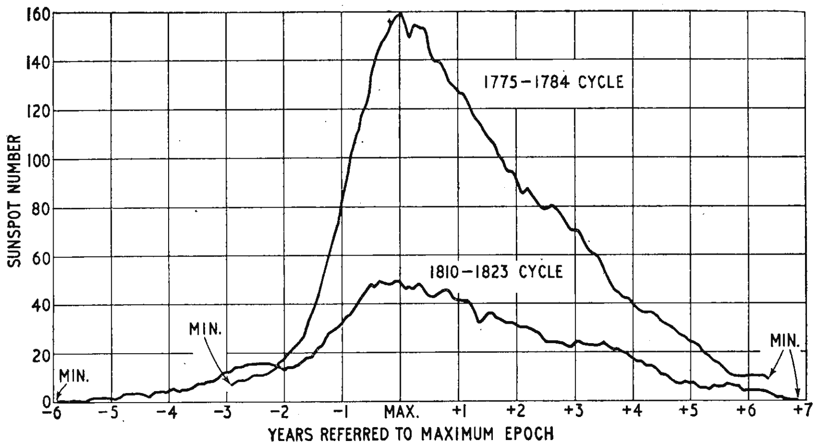


Fig. 2. The highest and lowest sunspot maximum in the records of the completed cycles.

If, in the northern hemisphere of the sun, the leader spot is of north and the follower of south magnetic polarity, then in the southern solar hemisphere the position is reversed, the leader spot being of south and the follower of north magnetic polarity. And this rule applies to all the bi-polar spots which are observed during the course of any solar cycle. But when the cycle is finished and the first bi-polar spots of the new cycle are observed a complete reversal of the magnetic polarities in both hemispheres has taken place, the leader being of south and the follower of north magnetic polarity in the northern hemisphere, and the leader of north and the follower of south magnetic polarity in the southern hemisphere. This offers another means of identifying spots belonging to a new cycle, and assists in the prediction of the cycle behaviour near the minimum epoch. Incidentally, it means that (disregarding the overlapping period) a complete cycle of solar activity lasts on an average about 22, rather than about 11, years, for the original conditions are not restored until two cycles of variation from minimum through maximum to minimum have elapsed. However, we shall, in this article, continue to regard a solar cycle as being the variation which occurs from minimum, through maximum, to minimum again.

The Relative Sunspot Number

Sunspot activity is observed and recorded in different ways, but the best-known criterion of the activity is the Relative Sunspot Number. This is obtained from solar observations made at different observatories according to the formula of Wolf:

$$R = k(10\,G + N)$$

where R = relative sunspot number,
G = observed number of sunspot groups,
N = observed total number of sunspots, either singly or in groups,
k = constant depending on type of telescope used.

It will be seen that this formula permits of the observations made at different observatories being reduced to a common basis, and also that it gives greater weight to sunspot groups than to individual sunspots. These groups are usually, in fact, the most active areas so far as the ionizing radiation is concerned and this weighting has produced a criterion of the general solar activity which, when compared with various phenomena which are known to depend upon solar activity, has shown remarkably good agreement. The observations made at the different observatories are correlated by that at Zürich and issued in the form of the Zürich Sunspot Numbers. A Number is issued for every day but the monthly mean of the daily values is the most widely used criterion of the solar activity.

The monthly sunspot numbers have been recorded for many years, and have also been calculated from former observations, so that there is a continuous record of them from 1749 to the present day. The monthly numbers themselves fluctuate widely from month to month, and it is best, in order to examine the general variation in the solar activity, to smooth out the short-period fluctuations by taking a twelve-month running mean value, each mean value thus obtained applying to the epoch at the centre of the twelve-month period which contains the monthly sunspot numbers taken. On the basis of this smoothed sunspot number we can now examine some sunspot cycles.

Features of the Sunspot Cycles

Between the sunspot minimum of February/March 1755 and that of April/May 1954 there were eighteen complete sunspot cycles. The records also include part of a cycle which was in progress when they began in 1749 and, of course, the present cycle, which began in April/May 1954 and is still in progress. The mean duration of a cycle in the eighteen completed cycles is 11.1 years but this figure is not a very useful criterion, for the cycles, in fact, varied in duration from 9.1 to 13.6 years. In Fig. 1 are plots of the smoothed sunspot number during the longest cycle (that which began at the epoch September/October 1784) and the shortest cycle (that which began at the epoch May/June 1766) which the records contain. Sunspot number is, in both cases, plotted against time from the preceding minimum. It is seen that though the time which elapsed between the minimum and maximum was exactly the same in both cases, the duration of the "declining" phase was well over four years longer in the case of the 1784-1798 cycle than in that of the 1766-1775 cycle. It should already be obvious, from these facts, that the prediction of the sunspot number for a long time ahead is likely to be a difficult and unreliable business.

The cycles also varied very considerably in regard to the highest degree of activity that was reached during their course. In Fig. 2 is plotted the smoothed sunspot number during the cycle with the highest maximum (that which began at the epoch June/July 1775) and that with the lowest maximum (that which began at the epoch June/July 1810) in the records of the eighteen completed cycles. It should be remarked here that the present sunspot cycle reached a maximum considerably higher than that of the 1775-1784 cycle, and is thus the cycle with the highest maximum ever recorded. As to the significance in the difference in the maximum sunspot number as between different cycles it should be remembered that the highest frequencies generally available for long-distance communication do not become usable until the smoothed sunspot number has a value of about 70, and thus, in a cycle like that of 1810-1823, they would not be usable throughout the entire cycle. Furthermore, there occurs one case in the records where the sunspot number did not reach a value near 70 during two *consecutive* cycles, which means that the higher frequencies would have been unusable for a period of some twenty-five years, as these cycles were both of longer duration than the average. It should also be borne in mind that "quasi-minimum" radio conditions prevail during the whole time the Number is below about 30, and this varies considerably according to the height of the maximum. During this period conditions for radio communication are particularly difficult, for only the lower frequencies are usable.

Characteristic Differences in the Cycles

It is evident therefore that, so far as radio communication is concerned, it is, at the beginning of a cycle, largely a matter of conjecture as to what the future holds in store, for both the duration of the coming cycle and the height of its maximum may vary very widely. However, in Fig. 2 the plots of sunspot number are made against time referred to the maximum epoch of both cycles, and one big difference in the two cycles is at once noticeable. Whereas, in the cycle with low maximum the cycle is roughly symmetrical about the maximum, having "increasing" and "declining" phases which are not very different in point of their duration, the cycle with high maximum is markedly "saw-tooth" in shape, having an "increasing" phase which has less than half the duration of the "declining" phase. In fact the records show that this is a general feature of sunspot cycles, i.e., the higher the maximum the shorter

TABLE 1

Type	Duration—Years		
	Min. to Max.	Max. to Min.	Whole Cycle
High Maxima ..	3.5	7.3	10.8
Medium-high Maxima ..	4.6	6.5	11.1
Low Maxima ..	5.8	5.7	11.5

is the "increasing" and the longer is the "declining" phase likely to be.

The eighteen completed cycles were therefore classified into three categories according to the height of the maxima, i.e., High Maxima where the sunspot number at maximum exceeded 116: Medium-high Maxima where the number at maximum was 80 to 116: Low Maxima where the number at maximum was below 80. There were eight cycles with High Maxima, four with Medium-high Maxima and six cycles with Low Maxima. In each category the mean value of the smoothed sunspot number for all the relevant cycles was taken for the maximum epoch, and for each monthly epoch from the maximum until the mean minimum epoch at the beginning and end of the cycles. The result is plotted in Fig. 3, the three curves of which may be taken as representing the mean sunspot number throughout cycles, respectively, with High, Medium-high and Low Maxima. Table 1 gives the mean duration times for the different phases of cycles in the three categories.

Thus the "mean" Low-Maxima cycle is almost symmetrical about the maximum, whereas the "mean" High-Maxima cycle has a "declining" phase of approximately twice the duration of the "increasing" phase.

The relationship between the sunspot number at maximum and the duration of the "increasing" phase is shown, for the 18 completed cycles, in Fig. 5, in which the straight line curve is the best fit for the plotted points, and indicates that the higher the maximum the shorter is the duration of the "increasing" phase. But Fig. 5 will not be of much use to us in the early part of a cycle for predicting what type of cycle it is likely to be, and one seeks for a practical method of doing this. It would appear that, in the early part of a cycle, the *rate* of

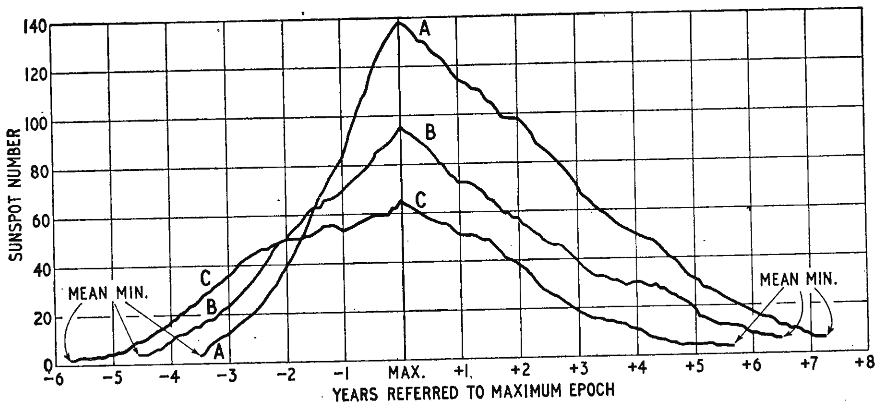


Fig. 3. Mean values of Sunspot Number: A, for cycles with High Maxima (8 cycles); B, for cycles with Medium-high Maxima (4 cycles); C, for cycles with Low Maxima (6 cycles).

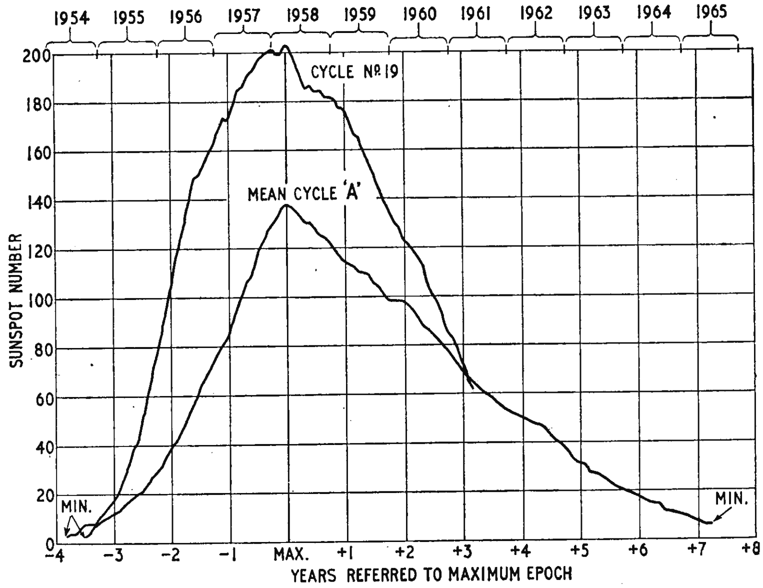


Fig. 4. Present sunspot cycle compared with the "mean" cycle with High Maximum

increase would be the best indicator of the height of the coming maximum and of the time when it is likely to occur. In Table 2 are given, for the first four six-monthly periods of the cycles shown in Fig. 3, the rate of increase in terms of sunspot numbers per month, and also the time which elapses before the maximum is reached. This shows that there is some relation between the early rate of increase and the height of the maximum, and also the time of its occurrence. In the last line of the Table similar data are given for the sunspot cycle now in progress, i.e., Cycle No. 19.

TABLE 2

Type	Increase per month—Sunspot Numbers				Time to Maximum Years
	First Six Months	Second Six Months	Third Six Months	Fourth Six Months	
High Maxima	1.4	1.7	2.8	4.0	3.5
Medium-high Maxima	0.6	1.1	1.2	1.8	4.6
Low Maxima	0.1	0.6	1.1	1.4	5.8
Cycle No. 19 ..	0.8	2.8	7.1	9.0	3.8

The Present Sunspot Cycle

During the first six months of the present cycle, therefore, the indication given by the rate of increase was that the cycle would be one with a Medium-high Maximum but, from then on, the rate of increase clearly indicated that it would be a cycle with a High Maximum, which indeed it was: the highest maximum yet recorded. The duration of the "increasing" phase was only slightly greater than that for the "mean" cycle with a High Maximum. In Fig. 4 are plotted the smoothed sunspot numbers for this "mean" High-Maximum cycle, and those for each month of Cycle No. 19 for which they the available, against time referred to the epoch of maximum. We may notice one rather striking thing about these two curves. Although the "increasing" phase of Cycle No. 19 was of relatively short duration, as would be expected for a cycle with a High Maximum, the behaviour during the "declining" phase has not indicated such a markedly saw-tooth shape as is usual in such a cycle. True, the decline has been considerably slower than the increase, but, since the first year after the maximum, it has been faster than that in the "mean" cycle. Thus, whereas the "mean" cycle showed decreases of 17%, 29% and 51% of the maximum value during each of the first three years following the maximum, the corresponding figures for Cycle No. 19 have been 13%, 40% and 65%. During the past two years, therefore, the sunspot activity has been decreasing at a more rapid rate than would have been expected and, up to the present, this rate shows little sign of decreasing.

The epoch of minimum sunspot activity from which Cycle No. 19 is assumed to take its start was at April/May 1954, and, if the cycle were to run the full course of 10.8 years which is the mean duration

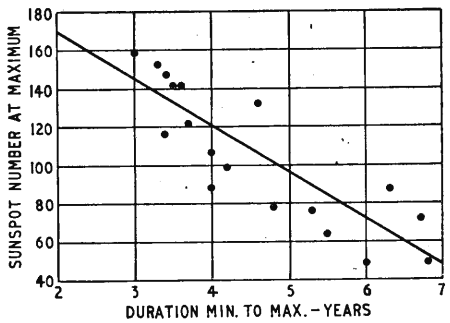


Fig. 5. Relationship between Sunspot Number at maximum and duration of period minimum to maximum.

of the former cycles with high maxima records, the next minimum epoch would be at February/March 1965. Since the increasing phase was a few months longer than that of the "mean" cycle it might have been expected that, if the markedly "saw-tooth" shape of a High Maximum were to be maintained, the next minimum might even be a few months later than that, say towards the end of 1965. But there will have to be a drastic change in the rate of decrease if the cycle is to last so long as that, i.e., if its "declining" phase is to have a duration of somewhat over 7.3 years. In fact, from the present rate of decrease—as seen in Fig. 4—it looks as if the minimum epoch might occur about two years earlier than had been expected, i.e., some time about the middle of 1963, in which case the total duration of the cycle would only be a little over 9 years. That, however, would be nothing very unusual, for three of the cycles with High Maxima previously recorded had total durations of under 10 years.

It would be as well, therefore, to be prepared for sunspot minimum as early as 1963 and for "quasi-minimum" radio conditions well before that. But—and this is intended to be the most emphatic statement in this article—no firm prediction of the time of the minimum epoch can, in fact, be made at this stage in the cycle. The rate of decrease in sunspot activity might well alter drastically: it might even increase for a time and so prolong the duration of the cycle considerably. Perhaps it will not be until the high-latitude spots of the new cycle appear that an indication of the time of minimum will be obtained. As to what may happen beyond the minimum, whether the next cycle will have a high or low maximum, how long the "quasi-minimum" radio conditions may last, these also are unanswerable questions at present. We cannot even begin to conjecture about them until the next cycle is well under way and we can see how the increase in activity shapes itself.

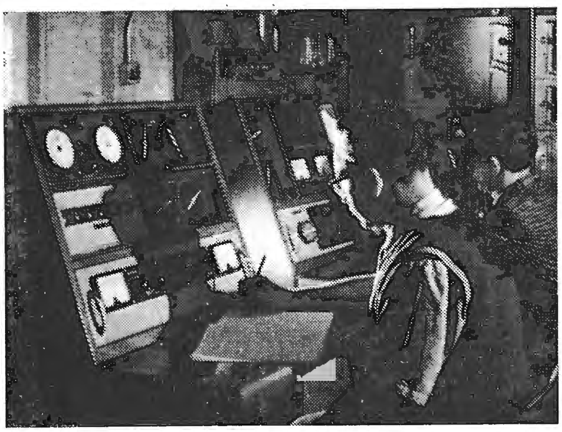
So with over 200 years of continuous sunspot records at our disposal we are unable to forecast with any accuracy how the activity will vary a few years hence, or exactly how long-distance radio conditions will be affected in that time. The sunspot records have been maintained for 200 years, yet from them we are unable to establish any law governing the sunspot cycles. But surely this is not surprising. For, after all, what is 200 years compared with the life of the sun?

METEOROLOGICAL RADAR

CONSTRUCTED on a unit system that allows the user to purchase only the parts of the equipment he needs for his particular purpose, the Cossor C.R.353 meteorological radar can be used for windfinding by tracking passive reflectors attached to balloons or for direct weather observation by the fitting of a suitable display unit. The C.R.353 can thus be used as a "minimum" equipment providing only one function or as a major component of a complete weather-reporting system.

The number of staff required for operation depends on the calls to be made upon the equipment; but even with the simplest assembly only two operators are required and, by the addition of automatic print-out machinery this can be reduced to one, or even none at all when the radar is fitted up for remote control.

Control console and (in background) part of transmitter and receiver. Operator is recording range, height and bearing of weather balloon.



The choice of 10cm wavelength for the transmitter allows the radar to operate "through" local rain.

Windfinding

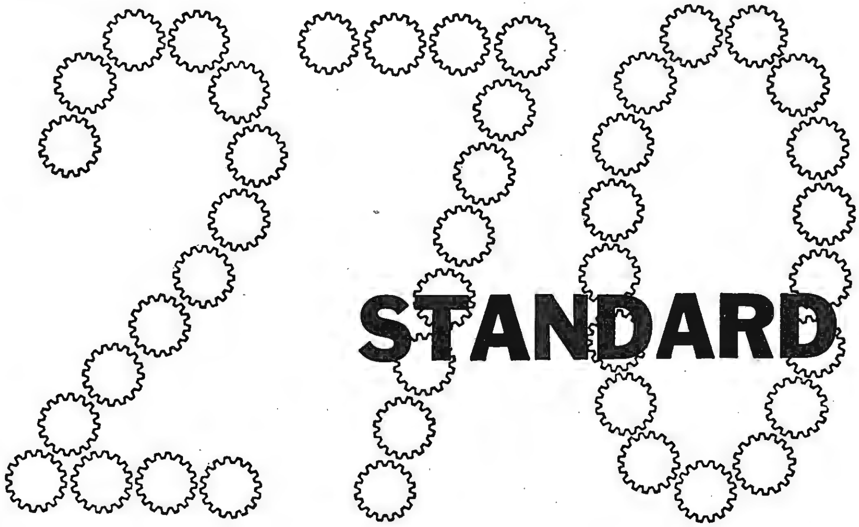
When the balloon is released it is sighted manually by a "putting-on" device to which the radar aerial is slaved. Once the target has been sighted by the radar it is tracked either manually by the operator, or automatically by the equipment. The three co-ordinates defining the position of the reflector are displayed on dial-type indicators on the Data Presentation Unit, the bearing and elevation indications being at intervals of 0.05°. The range indicator gives a six figure readout correct to the nearest 25m: maximum slant range in the windfinding role is 250km (120 nautical miles). A timing device is also incorporated in the display arrangement to provide pulses which freeze all the indicators at regular intervals so that the co-ordinates of the target, at the instant of each pulse, can be plotted or noted with ease.

Optional additions to the Data Presentation Unit can include an automatic print-out unit for the recording of position information and a computer to convert the slant range and elevation co-ordinates into ground range and height co-ordinates.

Cloud and Precipitation

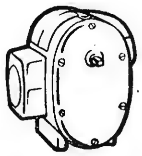
In this application the equipment is used to observe the position of storm centres and to assess densities of clouds and rain precipitation. The radar echoes are displayed on a plan position indicator with a 12-in c.r.t. This receiving unit incorporates iso-echo circuits which enable the operator to make cloud density determinations. Cloud heights can be estimated by varying the vertical tilt of the aerial.

Maximum range of the C.R.353 as a weather radar is principally determined by the earth's curvature and in practice is about 200 nautical miles. The displayed p.p.i. information can be conveyed by cable or radio link to a remote point and, if required, arrangements can be made for remote operation of the equipment.



STANDARD SPEEDS

from 57 mins. per rev. to 2,700 revs. per min.



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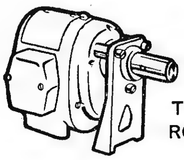

Type RQ 53

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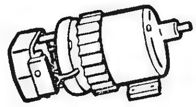

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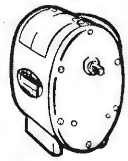

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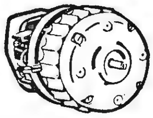

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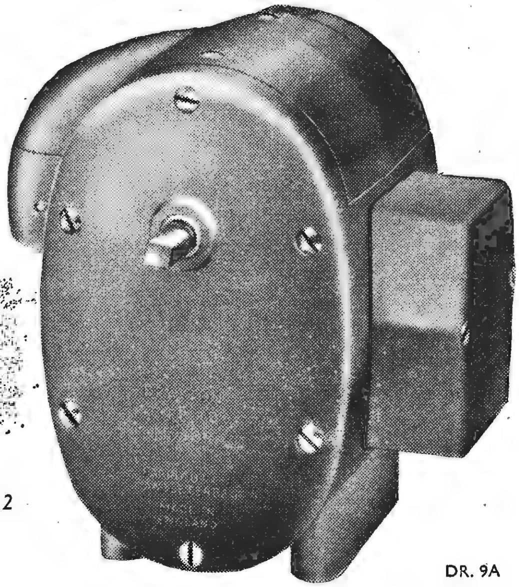

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DR. 9A

Transistor RC Oscillator

CURRENT DUAL OF THE WIEN NETWORK

By C. STOTT,* B.Sc., Grad. I.E.E.

IN phase-shift oscillators using transistors, a common practice seems to be to use the conventional phase-shift network of the old valve circuit and to design so that the transistor input impedance is high. A far more logical approach would be to exploit the transistor characteristic and to change the phase shift network to suit it. In this article the Principle of Duality is employed in the design of such an oscillator and the theoretical expectations and practical results are compared. The variation of oscillator performance with temperature is shown graphically.

Current Dual of Wien Network

The conventional Wien network is shown in Fig. 1 and is suitable for applications where a low source impedance and a high load impedance are available. Although transistor circuits can be designed to fulfil these requirements, it is preferable to use the inherent transistor properties of low input impedance and high output impedance. To do this, the Wien network of Fig. 1 must be rearranged to form its current dual. The requirement for such a current dual is that its transfer function is the same as its voltage counterpart, i.e.,

$$i_o = i_1 \frac{Z_2}{Z_1 + Z_2}$$

where Z_1, Z_2 are the same as those of Fig. 1.

The arrangement which gives this transfer function is shown in Fig. 2.

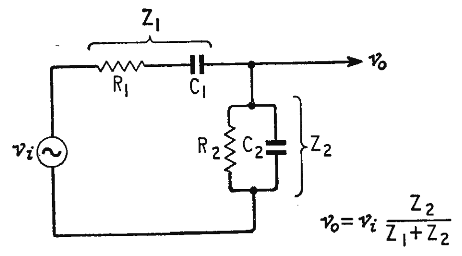

Fig. 1. Conventional Wien network

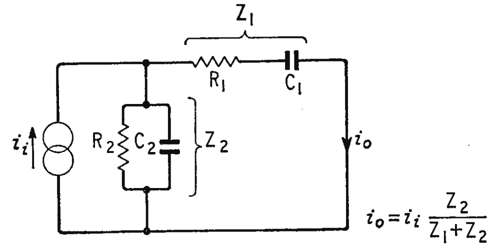

Fig. 2. Current dual of Fig. 1

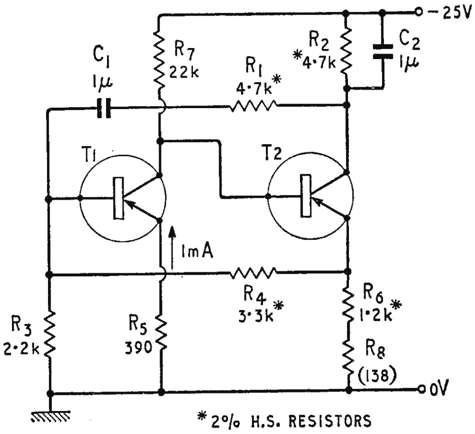

Fig. 3. Complete circuit of 34c/s oscillator

The conventional Wien network is thoroughly analysed in many standard text books† so time will not be taken here to repeat the analysis. However, the two principal results of this analysis are quoted in equations (1) and (2) and are equally applicable to the current dual considered here. It can be shown that the frequency of zero phase shift is

$$f_o = \frac{1}{2\pi\sqrt{C_1C_2R_1R_2}} \quad .. \quad .. \quad .. \quad .. \quad (1)$$

and that at this frequency, the current attenuation is

$$-A = \frac{i_o}{i_1} = 1 + \frac{R_1}{R_2} + \frac{C_2}{C_1} \quad .. \quad .. \quad .. \quad (2)$$

Equation (1) gives the frequency at which an oscillator employing this configuration will oscillate, and equation (2) gives the value of current gain needed to maintain oscillations.

Oscillator Configuration

By inspection of Fig. 2 and equation (2) the following list of requirements can be compiled:

(a) The network must be driven from a high impedance source.
(b) It must feed into a low impedance load.
(c) Somewhere between the load and source, there must be a current gain of A and also an amplitude limiter.
(d) There must be zero phase shift between input and output.

The configuration of Fig. 3 fulfils all these requirements.

If the voltage gain of T1 is G, then if G is large

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*English Electric Aviation Ltd.

†For instance, *Junction Transistor Electronics*, by R. B. Hurley published by Chapman and Hall

it can be simply shown that the input impedance is approximately

$$R_{in} = \frac{R_4}{G} \quad \ldots \quad (3)$$

and it is shown in the appendix that the current gain from T1 base to T2 collector is

$$A \simeq 1 + \frac{R_4}{R_6} \quad \ldots \quad (4)$$

The impedance looking out of the collector of T2 is considerably increased by the presence of the feedback resistor R_4 and is of the order of 1 or 2 megohms making T2 a good current source. For T1 to be an ideal current load its input impedance must be zero, but this is impossible to achieve in this simple circuit and a low value of impedance must be accepted. This can be achieved by making R_4 small and G large.

The action of the circuit is as follows:

A current $I\sin\omega t$ flows from T2 collector into the Wien network and if $\omega = \omega_o$, the output current is $-\frac{I}{A}\sin\omega_o t$. This flows to the virtual earth point at the junction of R_3 and R_4 and undergoes a current gain of $(1+R_4/R_6)$ before emerging again at T2 collector as $(1+R_4/R_6)\ I\sin\omega_o t/-A$. Thus provided that $(1+R_4/R_6) > -A$, oscillations will build up because the loop gain is greater than unity. Amplitude limiting is achieved by T1 just entering saturation and the forward current gain must be sufficient to ensure that it does this. When saturation occurs T1 ceases to amplify and the base and collector waveforms are then in phase, causing the Wien feedback to change sign and reduce the total loop gain. An equilibrium state is set up where the average loop gain is unity and constant amplitude oscillations ensue. This means that theoretically the output waveform will not be a pure sine wave, but in practice quite low harmonic distortion figures are found and are given later.

An important consideration in this type of limiter is temperature stability and fortunately this is achieved in the process of making T1 base a low impedance point. Leakage current flows from T1 base into this impedance and causes a temperature-dependent voltage to appear at the input. This voltage is amplified by the d.c. gain G of T1 and may cause T1 collector to enter saturation, making the output amplitude decrease to zero.

Design Procedure

A convenient supply of -25V was available and therefore the oscillator was designed for this voltage.

T1 and T2 were chosen to be BCZ11 transistors because of their low leakage currents and high β ($\beta = 30$).

(i) Choose V_{b1} to be low so that T1 can have high d.c. gain. Set $V_{b1} = -1$ volt.

(ii) This fixes V_{e1} to be approx. -0.4V as V_{be} is of the order of 0.6V for a silicon transistor in normal conduction. R_5 can now be chosen for a given collector current in T1. If this current is made say 1mA, then

$$R_5 = \frac{0.4}{1} = 400\,\Omega \simeq 390\,\Omega.$$

(iii) We now have to fix a d.c. level for the collector of T1 and it is this level which will determine the output amplitude of the oscillations.

A convenient value for R_7 is 22kΩ so that V_{c1} is -3 volts, allowing a 2.6 volt excursion before T1 bottoms. This swing is multiplied by the voltage gain of T2 to give the output voltage.

(iv) The total collector load of T1 is R_c and is composed of the 22kΩ resistor R_7 in parallel with the impedance looking into the base of T2. This impedance is

$$R_L = \beta\left[\frac{R_4\,R_6}{R_4 + R_6}\right] \text{ohms} \quad \ldots \quad (5)$$

provided we assume that the junction of T1 base and R_4 is a virtual earth.

$$\text{Now } \frac{R_4 + R_6}{R_6} \simeq 3.5$$

$$\therefore R_L \simeq \beta\frac{R_4}{3.5} \quad \ldots \quad (6)$$

Now if R_4 is large the loading on the collector of T1 will be light and the gain will be well defined by R_4 and R_6; however, the input impedance will be high. The requirements are for a well defined gain and for a low input impedance and a compromise value of R_4 was chosen to be 3.3kΩ.

From equations (5) and (6) R_L is approximately 30kΩ making $R_c \simeq 13$kΩ. The voltage gain of T1 is

$$G = \frac{R_c}{R_s + r_e + r_b\,(1-\alpha)}$$

where r_e is the internal emitter resistance of T1 and r_b is the base resistance.

For the transistor type used running at 1mA emitter current, $r_e + r_b\ (1-\alpha)$ can be found from the characteristic curves to be 40Ω.

$$\therefore G = \frac{13}{0.39 + 0.04} = 30.$$

The amplifier input impedance is from equation (3).

$$R_{in} = \frac{3300}{30} \text{ ohms} = 110 \text{ ohms.}$$

Because $V_{c1} = -3$V, V_{e2} sits at -2.4V and 1.4 volts are dropped across R_4. Thus a standing

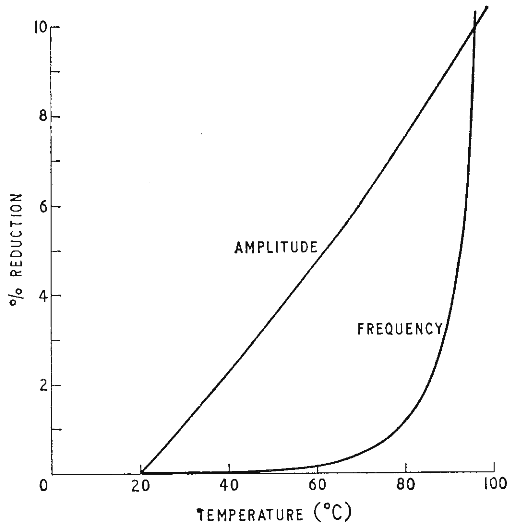


Fig. 4. Variation of amplitude and frequency with temperature

current of 0.425mA is drawn through it and is supplied by R_3 which is therefore 2.2kΩ. From these figures it is seen that the actual current supplied by R_3 is 0.455mA which is greater than that demanded by R_4. The result of this is that V_{b1} will sit at some potential a little more positive than − 1V causing the other d.c. levels in the circuit to adjust slightly to an equilibrium position.

(v) If the Wien network components are chosen so that $C_1 = C_2$ and $R_1 = R_2$, then from equation (2) the current attenuation $-A = 3$. Therefore $R_4/R_6 \geqslant 2$ to provide the required current gain. In the condition $R_4/R_6 > 2$, a small resistor may be connected in series with R_6 to adjust the current gain to the correct value without seriously affecting the d.c. levels in the circuit. With this type of gain adjustment, the oscillator may be used down to frequencies of fractions of a cycle/sec.

R_4 has been chosen to be 3.3kΩ.

$\therefore R_6 = 1.65$kΩ.

By making $R_6 = 1.2$kΩ the output stage runs at 2.4mA and we have some current gain in hand.

(vi) The only thing left to determine as far as the amplifier is concerned is the value of R_2 which is also one of the Wien network resistors. For T2 collector to sit half way between its top and bottom limits, V_{c2} should be − 13.7V. By choosing $R_2 = 4.7$kΩ then at 2.4mA, V_{c2} is exactly − 13.7V which is the optimum value.

(vii) We have previously stated that $R_1 = R_2$ $\therefore R_1 = 4.7$kΩ. The oscillator frequency is therefore determined by the value of C_1 and C_2 which again, are chosen to be equal.

From equation 1

$$\omega_o = \frac{1}{CR}$$

$$\therefore C = \frac{1}{\omega_o R} = \frac{10^6}{2\pi f_o \times 4.7 \times 10^3}\,\mu F$$

$$= \frac{33.8}{f_o}\,\mu F$$

By choosing $C_1 = C_2 = 1\mu F$ the frequency should be 33.8c/s and this frequency is used for the practical tests.

The design is now complete save for the determination of the gain adjusting resistor R_8 which will be of the order of a few hundred ohms and will be connected to earth in series with R_6. The inclusion of R_8 will not affect V_{e2} but will affect to a small extent V_{c2} by altering the available collector current. The d.c. loop will therefore be unaffected.

If the oscillator is to be used at elevated temperatures it is advisable to make the Wien resistors R_1 and R_2 high stability types in order to make the frequency dependence on temperature small. Also if R_4, R_6 and R_8 are made high stability then the amplitude stability will be good. The change in amplitude is due only partially to a change in forward current gain of the amplifier which ideally should remain constant. The main cause is a change in the values of the base-emitter voltages of the two transistors which have a marked effect on the collector voltage of T1 and therefore a proportional effect on the output amplitude.

Choice of R_8

This is determined experimentally by connecting a 250Ω potentiometer in series with R_6 and earth. With the potentiometer (R_8) at its minimum value the output waveshape should be seen to be distorting. The value of R_8 should be gradually increased until the distortion just vanishes and this represents the condition of maximum tolerable bottoming of T1. Further increase of R_8 should produce no change in output amplitude until the loop gain becomes too small, when the oscillations decay rapidly to zero.

R_8 should be adjusted so that T1 bottoms and the output amplitude is on the " flat " part of its characteristic. The potentiometer may now be replaced by resistors made up to the same value. In the present oscillator a value of 140Ω was required and was made up of a 120Ω and an 18Ω resistor.

Experimental Results

(a) Several values of C were connected in the circuit and the output frequency was measured by comparison with a decade oscillator using the Lissajous method. A short table of results is given below:

Capacitance (μF)	Calculated Frequency (c/s)	Measured Frequency (c/s)
50	0.686	0.7
1	33.8	33.7
0.25	135	135
0.1	338	340
0.02	1,690	1,690

It is seen from this table that the oscillator frequency can be accurately determined by calculation at the design stage.

(b) With the 1μF capacitors connected, the oscillator was operated at temperatures up to 95°C and changes in amplitude and frequency were noted. The results are plotted graphically in Fig. 4.

(c) The total harmonic distortion was measured for various temperatures up to 95°C. The maximum total distortion found was less than 6% and at temperatures up to 80°C the distortion was less than 4%.

(d) The supply voltage was varied ±5V on the nominal 25V with no change of frequency but the output voltage changed between +25% and −60%. This shows that the amplitude of the output is very much dependent on supply voltage as would be expected due to the nature of the limiter employed.

Improvement to Basic Design

If Fig. 4 is referred to, it can be seen that it is unwise to operate this oscillator at temperatures higher than 80°C otherwise the change in frequency may become intolerable. Also at this temperature, the amplitude has decreased by 7½% and this may also prove intolerable in more critical applications. The amplitude may be made more stable by cancelling the temperature drift of base/emitter voltages using a temperature sensitive element to stabilize the d.c. levels.

First let us examine the requirements of such a device with respect to Fig. 3. Suppose that i_{b1}

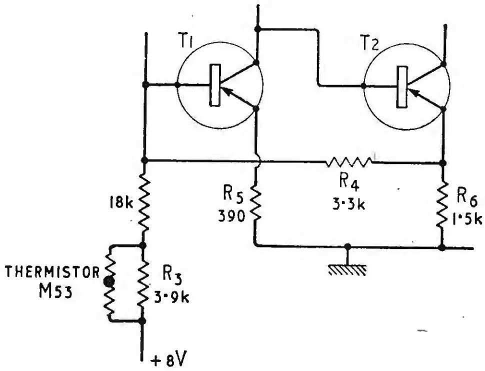


Fig. 5. D.c.-compensated circuit

is negligible and that V_{b1} tends to go more positive as the temperature increases, then the current through R_3 decreases making V_{e2} go more positive by an amount

$$V_{e2} = V_{b1}\left[1 + \frac{R_4}{R_3}\right]$$

The ratio R_4/R_3 plays an important part in the voltage drift so the first improvement is to increase R_3. As the rest of the circuitry is unchanged, the new R_3 must be taken to some positive voltage so that the same current is supplied through it. By including a diode in series with R3, the reduction in the forward voltage drop with temperature would tend to increase the current through R_3 and therefore oppose the effect of base-emitter voltage change. This method was tried successfully using three OA95 diodes in series with a 10kΩ resistor and resulted in a constant amplitude output at temperatures up to 65°C with some deterioration at temperatures above this. However, this method was considered to be expensive in diodes so a thermistor was used instead as a compensating element. This has the same action as the diodes and has the advantage of being more repeatable. A satisfactory arrangement is shown in Fig. 5 and with this arrangement the amplitude does not deviate from its value at 20°C by more than 2% up to a temperature of 80°C. R_6 may be increased to 1.5kΩ now that the shunting effect of the 2.2kΩ resistor has been removed. Distortion figures were again found to be less than 4% total harmonic up to 80°C.

Conclusions

A step by step method has been presented for designing a low frequency Wien oscillator with good temperature characteristics. The table shows that the design criteria, though simple and sometimes approximate are quite valid in practice and will enable an oscillator of a particular frequency to be accurately designed.

It might be pertinent to point out at this point that the 2% resistors specified for parts of this circuit are subject to a long-term drift of ±7% and for operation over a very long period, 0.1% resistors should be specified for the critical places.

It should also be mentioned that because of its lower frequency/phase slope, this particular oscillator will not exhibit as good a frequency stability as the Wien Bridge type of oscillator.

APPENDIX

Assume that T1 base is a virtual earth, and that R_c is unloaded by T2. The change of T1 collector voltage and therefore T2 emitter current is $V_{c1} = \alpha' i_b R_c$

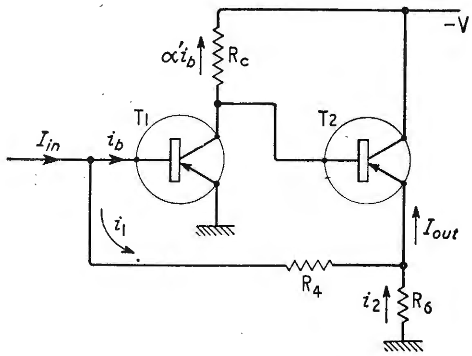
$$\therefore i_1 = \frac{i_b \alpha' R_c}{R_4}$$

$$\text{and } i_2 = \frac{i_b \alpha' R_c}{R_6}$$

$$\text{Now } I_{out} = i_2 + i_1 = \frac{i_b \alpha' R_c}{R_6} + \frac{i_b \alpha' R_c}{R_4}$$

$$= i_b \alpha' R_c\left(\frac{R_4 + R_6}{R_4 R_6}\right)$$

$$\text{and } I_{in} = i_1 + i_b = \frac{i_b \alpha' R_c}{R_4} + i_b = i_b\left(1 + \frac{\alpha' R_c}{R_4}\right)$$



$$\text{Current gain} = \frac{I_{out}}{I_{in}} = \frac{i_b \alpha' R_c\left(\frac{R_4 + R_6}{R_4 R_6}\right)}{i_b\left(1 + \frac{\alpha' R_c}{R_4}\right)}$$

if α' is large, this is approximately

$$\frac{I_{out}}{I_{in}} \simeq \frac{\alpha' R_c\left(\frac{R_4 + R_6}{R_4 R_6}\right)}{\frac{\alpha' R_c}{R_4}}$$

$$= 1 + \frac{R_4}{R_6}$$

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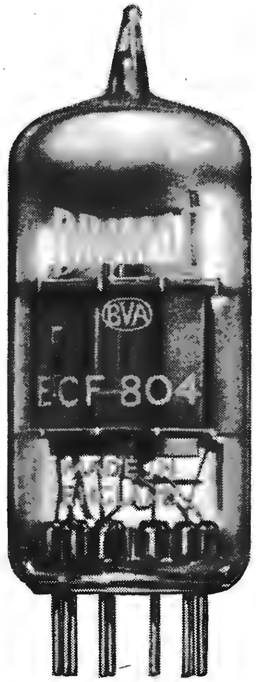

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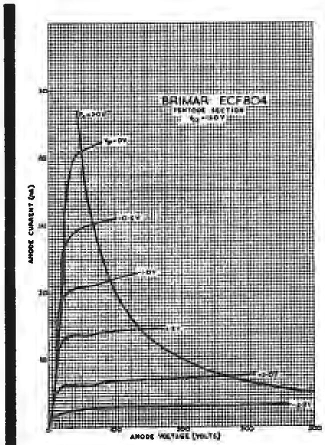

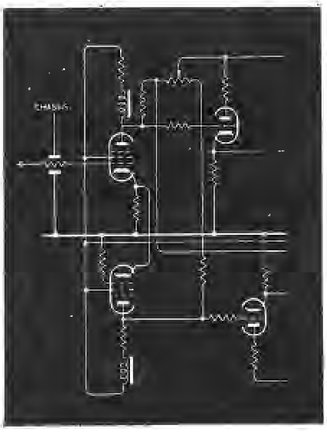

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TECHNICAL NOTEBOOK

"Zebra" Colour Television display tube was demonstrated recently at Sylvania-Thorn Colour Television Laboratories. Zebra is of the single-gun beam-indexing type and bears on its screen vertical stripes of phosphor in the three primary colours. On the back of the screen are indexing stripes of an ultra-violet emitting phosphor: light from these is picked up by a photomultiplier and used to time the conversion of the simultaneous N.T.S.C. information to a time-sequence of primary colour signals. At its present (experimental) stage of development the tube has about 1,100 phosphor lines on its face, so that, to achieve maximum purity of colour, a very fine spot has to be used. To avoid too severe a loss in brightness due to the small area of phosphor illuminated and a short life due to very heavy cathode loading, an elongated spot, formed by an elongated aperture in the gun, is used, with its major axis in the vertical direction. This, of course, also helps to reduce the visibility of the scanning lines (and could well be applied to black-and-white tubes). Zebra's versatility was demonstrated by plugging it into a black-and-white receiver, where it produced a satisfactory picture and its major advantage is that, in production, it should be little more expensive than a black-and-white tube.

Ultrasonic Amplification reported by Drs. A. R. Hutson, J. H. McFee and D. L. White of the Bell Telephone Laboratories in the Sept. 15, 1961, issue of *Phys. Rev. Letters* was obtained by applying a direct electric field (in the direction of propagation) to a piezoelectric semiconductor crystal. The gain depends both on the applied electric field and on the conductivity of the material. The latter was adjusted by altering the illumination, taking advantage of the fact that cadmium sulphide—the piezoelectric semiconductor actually used—is photoconductive. In a 7mm length of this substance gains of 18dB at 15Mc/s and 38dB at 45Mc/s were obtained. Quartz transducers were used to convert between the electrical input and output and ultrasonic vibrations in the crystal. These vibrations, because CdS is piezoelectric, produce an associated travelling longitudinal electric field: this in turn, because CdS is also conductive, causes electrons to flow in the crystal. The periodic piezoelectric field bunches these electrons and the direct electric field causes these bunched carriers to move in the direction of propagation faster than the ultrasonic vibrations. This results in amplification of these vibrations in much the same way as electromagnetic waves are amplified in a travelling-wave tube.

Flow Velocity Measurement of river water using ultrasonic equipment is described by H. F. Messias in *Electronics* for October 13, 1961. Two pairs of transmitting and receiving transducers are set up facing each other on opposite banks of the river, one pair being downstream from the other so that the signal path is at an angle to the water flow. A pulse-modulated carrier transmitted by the upstream transducer is received downstream in a time which is equal to $d/(v_s+v_\omega \cos\theta)$, where d is the signal path, v_s is the velocity of pressure waves in water, and v_ω is the water velocity. The received pulse is used to re-trigger the upstream modulator, the result being a train of pulses with a time between pulses equal to the transit time of the pressure waves. The same thing happens to the signal from the downstream transmitter, except that the water velocity component is negative and the transit time is increased. The two signals are detected, the p.r.f.'s multiplied to convenient frequencies and used to drive a pair of synchronous motors, which in turn drive a differential tacho-generator unit producing a d.c. output proportional to half the difference frequency of the two inputs. After allowance for the angle between water flow and transmission path, and setting-up of transducer separation, the output of the tacho-generator is used to drive an indicator calibrated in feet per second.

New Microwave Power Amplifier suggested by P. A. Clavier in a letter in the June 1961 issue of *Proc.I.R.E.* (p. 1083) would consist simply of a Lecher line or coaxial cable in which one of the conductors (the cathode) is made electron emissive. A direct voltage is applied between the cathode and the other conductor (anode) in such a direction as to repel electrons towards the cathode. If a sufficiently large signal then travels down the guide, the appropriate peaks of this signal will counteract the repulsion voltage and allow electrons emitted from the cathode to reach the anode. It can be shown that the anode current so produced contains an a.c. component in phase with the applied signal. Theoretically, efficiencies as high as 60% for power gains of 50 should be achievable with this type of device. Essentially it can be considered as a microwave class-C amplifier. The series of harmonics thus produced may be of value: alternatively the energy can be confined to the fundamental by using a resonant load.

"Solid Circuit" Computer was recently demonstrated by Texas Instruments in conjunction with the U.S. Air Force. Occupying only 6.3 cu in and weighing only 10 oz, it is functionally similar to a computer using 8500 conventional components and with 150 times the volume and 50 times the weight. The new computer uses 587 of the Texas Series 51 digital silicon Solid Circuits. Each of these is formed from a standard silicon bar and hermetically sealed in a package only $\frac{1}{4}$ by $\frac{1}{8}$ by $\frac{1}{32}$ in. Six different types of circuit are available and all can operate from -55 to $+125°$C. These circuits need so little power that the whole computer consumes only sixteen watts. Although normally a general-purpose device, for the demonstration it was programmed as a desk calculator capable of adding, subtracting, multiplying and taking square roots. It has an operand word length of 10 bits plus sign.

"Colourless" Artificial Reverberation delay system developed at the Bell Telephone Laboratories is described by M. R. Schroeder and B. F. Logan in an article under this title in the July 1961 issue of the *Journal of the Audio Engineering Society.* In the normal type of artificial reverberation delay system the signal is fed through a circuit producing a fixed time delay and at the same time the output of the delay circuit is amplified (by a factor g say) and fed back to its input. This normal system has however two disadvantages: its amplitude/frequency response is comb-like (with peaks at integral multiples of the reciprocal of the delay time) and it also produces many fewer echoes than a normal room. The former disadvantage results in a "hollow" sound and the latter in "flutter" effects with transient sounds. These disadvantages are avoided in the Bell system: a flat amplitude/frequency response is obtained by multiplying the output from such a normal system by $1-g^2$ and adding it to the undelayed signal multiplied by $-g$; and the number of echoes is increased sufficiently by connecting in series five such flat response systems having incommensurate delays.

News from Industry

Elliott-Automation has acquired from Firth Cleveland the capital of Firth Cleveland Instruments, which will continue to operate from its premises at Treforest, Glamorgan. Its name is being changed to Elliott (Treforest) Ltd. Firth Cleveland states that the net effect of the transaction will be to improve group liquidity by some £350,000, and it is consistent with the group's policy of reshaping its commitment in the instrumentation field as evidenced by the disposal last October of the majority holding in the Solartron Electronic Group, which was acquired by Schlumberger of Houston, Texas.

U.K./Czechoslovak Trade.—Quotas for Anglo-Czechoslovak trade in 1962 announced recently by the Board of Trade make provision for the exporting of U.K. measuring instruments to the value of £11,000 and electronic and communication equipment, including domestic receivers, sound reproducing equipment, valves and components, to the value of £110,000. Czech goods for import to U.K. include valves and components for sound radio and TV sets to the value of £80,000 (of which not more than £12,000 is for transistors).

U.K. Quotas for Japanese Imports.—Under new trade arrangements between the U.K. and Japan, the Board of Trade have announced the following import quotas for the period to September 30 next: transistor and other than transistor gramophones and radio gramophones (£150,000 and £300,000 respectively); telecommunication cables (£24,000); transistor (£500,000) and valve (£300,000) sound radio and TV sets; basic parts and accessories for radio apparatus other than semiconductors (£200,000); semiconductors (£150,000); and scientific and precision instruments other than optical instruments (£70,000).

BFN "One Man" Studio.—A new studio control room, recently installed by Telefunken G.m.b.H. at the British Forces Network station in Berlin-Spandau, enables the compere to combine, single-handed, B.B.C. overseas services, disc and tape recordings and live local commentaries into a popular programme which has a large following, not only among Services personnel but also the local population. The picture shows the duplicated tape machines, and between them the mixing controls and the announcer's microphone.

Coin-in-the-Slot service depots for marine radio and electronic equipment, devised by the Marine Division of Pye Ltd., are being operated by J. Davy Ltd. in Kent as an aid to yachtsmen sailing the south-east coast. The new system allows owners of Pye marine equipment to leave any unit due for servicing in a special theft-proof and weather-tight locker installed at the harbourside, and later collect the serviced equipment from the locker. Keys are provided to users, and a small coin-in-the-slot fee is charged for use of a locker. Service charges are settled in the normal way. The first lockers are being installed at Ramsgate.

Industrial Research.—Results of a survey, including commentaries and a statistical report, on "Industrial Research in Manufacturing Industry 1959/60," are now available from the Federation of British Industries, 21 Tothill Street, London, S.W.1. Prepared by the F.B.I. in conjunction with the National Institute of Economic & Social Research, the survey is described as the first investigation to attempt a qualitative as well as quantitative analysis of research and development in British manufacturing industry.

Nigerian Electronics Ltd. has commenced production of sound radio and television receivers and air conditioning equipment at its newly built factory at Apapa, near Lagos. It is stated to be the first plant in West Africa to undertake such production. First product is a portable transoceanic radio, which is being manufactured under a technical-assistance licence granted by Westinghouse. This new venture follows research and planning by the Rockefeller Brothers Fund and development by the Auriema Group, consultants, of New York.

Radio Rentals Ltd.—Group trading profit for the year to August 31st, 1961, was higher at £6,497,337 as compared with £5,742,065 for 1959/60. The rise was offset by heavier depreciation and tax provisions, leaving the profit balance little changed at £1,570,401 (£1,573,987).

Contactor Switchgear.—Profit for the year to July 31st, last, after taxation, is up at £83,037 as compared with £58,209 for the preceding year. During the period under review the company acquired the share capital of Radio Aids, Ltd., which company has now been renamed Contactor Switchgear (Electronics) Ltd.

Ronette Distribution.—H. K. Harrisson & Co. Ltd., 73 Great Titchfield Street, London, W.1, have been appointed sole distributors for Ronette microphones and pickups, which were previously handled by Trianon Electric Ltd.

Aero Electronics Ltd., of Gatwick House, Horley, Surrey, have been appointed sole U.K. distributors for the Data Systems Division of Harman-Kardon Inc., New York, producers of logic modules and digital instrumentation.

Dage (Great Britain) Ltd., of Rickmansworth, Herts, have recently been appointed U.K. distributors of equipment manufactured by Duncan Electronics Inc., of Costa Mesa, Cal., and Microwave & Semiconductor Instruments Inc., of Richmond Hill, New York. Duncans specialize in potentiometers.

Business Equipment Trade Association is the new title of the Office Appliance & Business Equipment Trades Association.

Britain's "teenage" market, found to be worth over £1,000M in 1959, is to be surveyed again this year. Covering 4½M unmarried people between the ages of 16 and 24, the survey will concentrate on a range of durable goods, including tape recorders, records, record reproducers, and sound radio receivers. The report will provide such information as ownership by make, type and whether new or secondhand; purchases during the past year; prices paid and method of payment. The Survey will be made by Market Investigations Ltd., of Berners Street, London, W.1.

S.G.S.-Fairchild Ltd. is the name of a new company being formed in the U.K. by Societa Generale Semiconduttori of Milan, to supply semiconductors. The Fairchild Camera and Instrument Corporation, of New York, has a substantial interest in S.G.S. of Italy, which is the licensee for Fairchild semiconductors, with exclusive distribution rights in Europe. The London offices of S.G.S.-Fairchild are at 30 Berkeley Square, W.1.

Radio Eireann have contracted Marconi's Wireless Telegraph Company for the supply and installation of a Band I television transmitting station at Maghera, Co. Clare, Ireland. The vision transmitter will be rated at 8kW, and it will operate on the 625-line system (8Mc/s channel). Its associated sound transmitter will have an output of 2kW.

Stock records of 90,000 different types of naval stores for H.M. ships, held at the Royal Naval Stores Depot at Copenacre, Wiltshire, will soon be maintained by an EMIDEC 1100 computer which was recently delivered to the Depot by EMI Electronics Ltd. The computer will also maintain lists of components for each of some 1,500 different types of electronic equipment.

The new Trident airliners are equipped with h.f. radio communication equipment, radio altimeters for automatic landing, an aircraft inter-communication system, and aerials by Standard Telephones & Cables Ltd. It is also announced that Ekco Electronics Ltd. has received an order for its latest Type E190 transistorized weather radar equipment for installation in the twenty-four de Havilland Trident aircraft recently ordered by British European Airways.

World-wide radiotelephone facilities are provided for passengers of the new "hotel" liner *Transvaal Castle,* recently handed over by John Brown & Co. (Clydebank) Ltd. to the Union Castle Steamship Co. Ltd. Two powerful Marconi Globespan radiotelegraph/telephone transmitters and two Atlanta receivers are installed in the new liner, which will be engaged on the South African Mail Service.

I.T.A. in N. Ireland.—Koram Hill, near Strabane, 20 miles south of Londonderry, is the site of a new 1,000ft mast which the I.T.A. are having built to extend their coverage to West Ulster. E.M.I. Electronics are supplying the mast and aerial, which will be vertically polarized and will transmit on channel 8. Sub-contractor for supply and erection of the mast is B.I.C.C. It is hoped the station will be completed by next summer.

Marine Radio Stations at Malin Head in the north and Valentia in the south west extremity of Ireland are to be re-equipped with a number of Pye 1kW m.f. transmitters. The transmitters are equipped for 3-channel operation and have remote control facilities.

"**Resista**" sub-miniature, high-stability resistors, marketed by GASP (G. A. Stanley Palmer & Co. Ltd.), Maxwell House, Arundel Street, London, W.C.2, now include a new Type Rsx1 composition film resistor.

Another tactical control radar, which continuously provides an accurate 3-dimensional picture of all aircraft in the battle zone with facilities for rapidly "locking on" to selected targets the precision target illuminating radar, has been handed over to the R.A.F. by Associated Electrical Industries Ltd., who, in conjunction with the Royal Radar Establishment, Malvern, was responsible for its design and manufacture.

The British Army is to adopt the Vickers Vigilant one-man anti-tank guided missile, and will use it both in the infantry role and by reconnaissance units of the Royal Armoured Corps. Developed entirely with private capital, the Vigilant will be put into quantity production at the Stevenage factory of British Aircraft Corporation.

Lunartron Electronics Ltd., of 42 Langley Street, Luton, Beds., have formed a Process and Component Heating Division, which undertakes to heat solids and liquids in connection with various industrial processes, and to provide heating elements, suitably insulated, for attachment to customers' assemblies.

Painton & Company, the Northampton component manufacturers, are undertaking a £500,000 development scheme which will provide additional manufacturing and office space on an eight-acre site adjoining their existing factory.

"CABMA Register 1961-2," appearing in its ninth edition, is the Canadian buyers' standard work of reference to British goods and makers. 620 pages. Price 15s 0d, post free, from Iliffe Books Ltd., Dorset House, Stamford Street, London, S.E.1.

Levell Electronics Ltd. have moved to new premises at The Trading Estate, Park Road, Barnet, Herts. (Tel.: Barnet 5028).

OVERSEAS TRADE

Ekco Export Ltd. has been formed as a subsidiary of E. K. Cole Ltd. to handle the export of sound radio, television and car radio receivers, electric heating and plastics products produced by the Ekco Group. T. C. Cleveland, formerly export manager of E. K. Cole Ltd., is director and general manager, the other directors being W. M. York (chairman) and J. Corbishley. The new company will operate from the main Ekco Works at Southend-on-Sea and a Personal Exports Department and showroom for sound radio and television receivers will be opened in London at 41-47 Old Street, E.C.1. The previous Export Sales Department of E. K. Cole Ltd. at 5 Vigo Street, London, W.1, has been transferred to Southend-on-Sea.

Two Pye Installations.—A Pye Instrument Landing System, incorporating a directional localizer, has been selected by the French Aviation Authorities for installation at their research and development centre at Bretigny, near Paris, where work on military and civil aircraft landing systems is carried out.

Cossor Equipment for Bahrein.—Cossor Communications Company have supplied a medium power v.h.f. radio station to the Port of Bahrein Authority, where it will be used to control marine traffic using the harbour. In addition, public correspondence channels provided, will enable ships' passengers and crews to converse with subscribers to the Bahrein public telephone system.

Swedish Order for Marconi.—The Royal Board of Swedish Telecommunications has placed another large order for television and sound broadcasting transmitters with Marconi's Wireless Telegraph Company. The new contract calls for the supply of 21 vision transmitters, 21 sound transmitters, 40 frequency-modulated sound transmitters and a considerable quantity of programme input, paralleling, feeder and ancillary equipment.

The Swedish Navy Board has placed a substantial order for radar display units with the Electronic Apparatus Division of Associated Electrical Industries Ltd. These units are to operate in conjunction with coastal surveillance radar equipment previously supplied by A.E.I.

Belgrade's new airport is being equipped with a complete airfield radar surveillance system made by Cossor Radar & Electronics Ltd., of Harlow, Essex. The equipment, known as CR787, weighs over 10 tons.

The University of Sydney has ordered a KDF9 electronic computing system from the English Electric Company, for its School of Physics.

Greenland's Narssarssuaq weather station, at the direction of the Danish Meteorological Institute, has recently completed the installation of Decca Type WF2 windfinding radar equipment.

FEBRUARY MEETINGS

Tickets are required for some meetings; readers are advised, therefore, to communicate with the secretary of the society concerned.

LONDON

7th. Women's Engineering Society.—"Space research and radio" by E. Golton at 7.0 at "Hope House," 45 Great Peter Street, Westminster, S.W.1.

8th. Radar and Electronics' Association.—"Colour television" by I. J. P. James at 7.0 at Royal Society of Arts, John Adam Street, Adelphi, W.C.2.

9th. I.E.E.—"The new transatlantic telephone cable" by R. J. Halsey at 5.30 at Savoy Place, W.C.2.

9th. Television Society.—"Problems associated with the B.B.C. 'Tonight' programme" by W. Cave at 7.0 at Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, W.C.2.

12th. I.E.E.—"Dilogarithms and associated functions" by L. Lewin at 5.30 at Savoy Place, W.C.2.

14th. I.E.E.—Faraday lecture "Expanding horizons in communications" by D. A. Barron at 5.30 at the Central Hall, Westminster, S.W.1.

14th. Brit.I.R.E.—"The application of rapid access photographic techniques to radar display" by S. R. Parsons at 6.0 at London School of Hygiene and Tropical Medicine, Keppel Street, W.C.1.

16th. I.E.E.—Discussion on "How should we teach microwaves?" opened by Prof. A. L. Cullen at 6.0 at Savoy Place, W.C.2.

16th. B.S.R.A.—"Sound recording in schools" by John Weston at 7.15 at Royal Society of Arts, John Adam Street, Adelphi, W.C.2.

19th. Royal Society of Arts.—"The influence of the industrial and technological professions on economic life" by A. M. Holbein at 6.0 at John Adam Street, Adelphi, W.C.2.

21st. I.E.E.—"Semiconductor static switching" by D. D. Jones at 5.30 at Savoy Place, W.C.2.

21st. Brit.I.R.E.—"Some factors which determine sensitivity of cells to ionizing radiation" by Dr. P. Alexander at 6.0 at London School of Hygiene and Tropical Medicine, Keppel Street, W.C.1.

23rd. Television Society.—"Television in flight simulators" by E. T. Emms at 7.0 at Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, W.C.2.

26th. Royal Society of Arts.—"The influence of the scientific professions on economic life" by Sir Patrick Linstead at 6.0 at John Adam Street, Adelphi, W.C.2.

27th. Society of Instrument Technology.—"On-line computers for process control" by J. F. Roth at 7.0 at Manson House, 26 Portland Place, W.1.

28th. I.E.E.—"The application of low noise reception techniques" by F. Graham Smith at 5.30 at Savoy Place, W.C.2.

28th. Brit.I.R.E.—"Stereo pickups" by S. Kelly at 6.0 at London School of Hygiene and Tropical Medicine, Keppel Street, W.C.1.

28th. British Kinematograph Society.—"Medium screen colour television projector" by Peter Lowrie at 7.30 at Colour Film Services, 22-25 Portman Close, Baker Street, W.1.

BELFAST

13th. I.E.E.—"The banana-tube display system—a new approach to display of colour television pictures" by Dr. P. Schagen at 6.30 in the Lecture Theatre, David Keir Building, Queen's University, Stranmillis Road.

BIRMINGHAM

26th. I.E.E.—"High quality loudspeakers" by D. E. L. Shorter at 6.0 at the James Watt Memorial Institute.

BOLTON

1st. Brit.I.R.E.—"Computers and data processing" by J. A. Bailey at 7.0 at Bolton Technical College.

BRISTOL

14th. Brit.I.R.E.—"The radio interferometer systems" by T. Carter at 7.0 at the University Engineering Lecture Rooms, Queens Building, University Walk.

21st. Society of Instrument Technology.—Discussion on "Instrument design" at 7.30 at Department of Physics, University of Bristol, The Royal Fort.

CARDIFF

7th. Brit.I.R.E.—"The function of a computer" by Dr. A. D. Booth at 6.30 at the Department of Physics, University College of South Wales.

CHESTER

22nd. Society of Instrument Technology.—"Electronic versus pneumatic control" by C. H. Gregory at 7.0 at Stanley Place, Watergate Street.

CHESTERFIELD
21st. I.E.E. Graduate & Student Section.—" Recent trends in loudspeaker design " by R. E. Cooke at 7.0 at the College of Technology.

DUBLIN
5th. I.E.E.—" Engineering aspects of Irish television " by T. Parker and J. H. Harbison at 6.0 in the Engineers Hall, Dawson Street. (Joint meeting with the Institution of Civil Engineers of Ireland.)
15th. I.E.E.—" The banana-tube display system—a new approach to display of colour television pictures " by Dr. P. Schagen at 6.0 in the Engineers' Hall, Dawson Street.

EDINBURGH
7th. Brit.I.R.E.—" Transistors in transmitters and communications receivers " by A. J. Rees at 7.0 at the Department of Natural Philosophy, The University, Drummond Street.

GLASGOW
8th. Brit.I.R.E.—" Transistors in transmitters and communications receivers " by A. J. Rees at 7.0 at the Institution of Engineers and Shipbuilders, 39 Elmbank Crescent.

LIVERPOOL
8th. Society of Instrument Technology.—" Modern trends in instrument installation " by A. J. Matthews at 7.0 at the Merseyside and North Wales Electricity Board Industrial Centre, Paradise Street.

MIDDLESBROUGH
7th. I.E.E.—" Silicon power rectifiers " by A. J. Blundell, A. E. Garside and R. G. Hibberd at 6.30 at the Cleveland Scientific and Technical Institution.
21st. Society of Instrument Technology.—" Training of instrument engineers " by Professor J. E. Parton at 7.30 at Cleveland Scientific and Technical Institution.

NEWCASTLE-UPON-TYNE
14th. Brit.I.R.E.—" Some physical and physiological signal-to-noise ratios " by Dr. L. Molyneux at 6.0 at the Institute of Mining and Mechanical Engineers, Neville Hall, Westgate Road.

NEWPORT
28th. Society of Instrument Technology.—" The training of instrument engineers " by J. C. Stone at 6.45 at the Newport and Monmouth College of Technology.

OXFORD
6th. Institution of Production Engineers.—" Electronic data processing " at 7.30 at the Town Hall.

SOUTHAMPTON
28th. Brit.I.R.E.—" Some aspects of the conversion of heat to electricity " by Dr. B. J. Hopkins at 7.0 at the Lanchester Building, The University.

TORQUAY
8th. I.E.E.—" Microminiaturization " by L. J. Ward at 3.0 at The Electric House, Union Street.

WOLVERHAMPTON
7th. Brit.I.R.E.—" Some aspects of industrial electronics " by F. C. Riches at 7.15 at Wolverhampton College of Technology.

UNBIASED

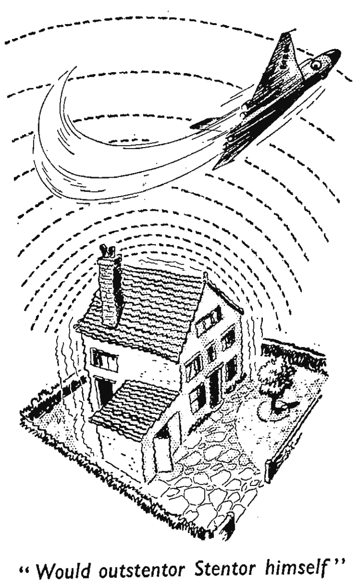
By "FREE GRID"

Jet-Controlled A.V.C.

RECENTLY I spent a few days with a friend who lives within a few miles of one of our big airports and he is in the approach area where the big transoceanic jets are flying fairly low.

I can only say I was truly appalled at the difficulties of listening to broadcasting amid the din of these super aircraft. On some days it is worse than others as the jets vary their direction according to the way the wind is blowing. The strangest thing of all, to my mind, is that the people most intimately concerned don't seem to bother much about it, although I notice they always seem relieved when the uproar has temporarily ceased. They remind me of the man who frequently hit himself on the head with a mallet as the feeling was so pleasant when he left off. No doubt our psychiatrists will be able to furnish a glib explanation of this, as they do of most other quirks of human nature.

When a jet approaches, it is necessary for listeners in the district to rise from their seats and turn the volume control to the "full-blast" position and lower it again after the plane has passed. Surely this could be done far better by a microphone on the roof or, better still, at some distance from the house to obviate feedback, which, by means of the necessary servo mechanism or electronic "circuitry," would regulate the volume of sound from the set in relation to the noise of the passing plane.



"*Would outstentor Stentor himself*"

The *modus operandi* of my idea would need carefully working out, for we don't need to increase the *sensitivity* of the set, and thereby bring up the background noise. It is merely the audio gain we want to adjust and so I think, therefore, that my title of "jet-controlled a.v.c." is correct.

There is a still more unfortunate class of listener, living on the very fringes of an airport, to whom turning the wick of their sets full up is quite inadequate. To them I would suggest the installation of a powerful p.a. system which the roof microphone could switch into circuit when necessary, so that the voice of the announcer would outstentor Stentor himself, and the most pianissimo passages of music be raised to the level of Wagner at his wildest.

We don't want a lot of scrappy bits and pieces added to sets, and so I hope our big manufacturers will be willing to turn out special sets fitted with the necessary apparatus for sale in areas near our great airports. What is wanted is a set with a jack into which it is only necessary to plug the roof microphone.

Pipeless P.A. Pioneers

LAST November I wrote in these columns about a first class entertainment I had seen on the pier of a well-known seaside resort. I also expressed the opinion that it was marred by the trailing cable connecting singers and others to the p.a. system and I suggested that this be replaced by a microwave link.

I did not mention the name of the resort but spoke of it in such glowing terms that the entertainments manager of Southend-on-Sea, Essex, wrote to the editor saying, quite rightly, that he felt sure I must have been talking about the pierhead show there. He also asked for details of the cableless link which I advocated, as he felt it would fill a need in certain of the resort's entertainments in the coming summer.

The editor gave him the required information together with the names of some firms who make this equipment commercially. It may be, therefore, that in the approaching summer Southend will find itself hailed as the pioneer of pipeless p.a. among the seaside resorts of this country, but it will certainly not be the pioneer as far as other countries are concerned. A Singapore reader has written to tell me that this system is in use in various places of entertainment there, the apparatus being of Japanese manufacture.

The midget Japanese transmitter measures 5in × 2½in × 1in and so is easily held in the singer's hand. It is fully transistorized and works on 27Mc/s, feeding 60 milliwatts into an aerial of fine wire sewn into the gown of a lady singer, presumably a male singer would have it sewn into the seat of his trousers.

My Singapore correspondent does not give me details of the dimensions of the aerial, but maybe a professional singer who would use the apparatus frequently might have the antenna tatooed on her in conductive flesh coloured ink; in fact the printed circuit itself could be tattooed on the skin in the manner I suggested for the "Torso Two" personal receiver in these columns some time ago. It would be necessary, of course, to embed the tiny transistors under the skin by means of plastic surgery.

However, in all seriousness, I do hope to find a "pipeless p.a." link in use at Southend next season. If so, I will certainly let you know of it through these columns.

Debawling Device

IN the July issue of the *Post Office Magazine*, I read an interesting account of two special types of 'phone which the P.M.G. has made available for the benefit of those who have the misfortune to be aurally or vocally defective.

First there is a 'phone for those who are hard of hearing. In this instrument, a miniaturized amplifier is built into the handset to boost incoming signals. The second instrument employs a rather larger amplifier in a separate case which is intended to be screwed to the skirting board. This device amplifies the voice of the person speaking into the microphone.

While I admire the ingenuity of the Post Office engineers in producing these instruments, it astonishes me that they have overlooked the necessity of producing another special type of telephone which, I think, all of us need from time to time. Every 'phone user knows the type of person who bawls with ear-splitting intensity into the microphone. Such people are presumably under the impression that the telephone is a speaking-tube rather than an electrical instrument, although even with a speaking tube, bawling does not really help.

What we need is that every 'phone be fitted with a simple debawling device in the earpiece circuit so that we can listen in comfort to those who think they are on the parade ground when using the 'phone. All that is needed is a push button to bring into circuit a simple fixed attenuator when necessary.