Wireless World

NOVEMBER 1954 VOL. 60 No. 11

B.B.C. Report

THE recently published Report* to the Postmaster-General for the year 1953-54 by the B.B.C. Board of Governors loses a little of its interest and value through being a trifle out of date. The document was prepared just before the final passing of the Act setting up the Independent Television Authority, and so, naturally enough, any attempt to consider the position of the Corporation *vis-à-vis* the new competitive body is specifically disclaimed.

Such incidental mention as there is of the possible effects of competition is confined to finance. It is a matter of some concern that, on present estimates, the B.B.C. will receive about £6M less than it expects to need for carrying through development plans during the next three years. Even if the financial stresses of that period can be weathered, the Corporation will, in the words of the Report, "inevitably have to ask for a larger share or even the whole of the proceeds of the £1 and £3 licences thereafter." We imagine that public opinion, and especially opinion in the radio world, will be strongly in favour of allowing the B.B.C. sufficient finance to carry through at least those parts of its development plan that have already been approved in principle by the Government. British broadcasting has been built up on a basis of successful long-term planning and even its harshest critics will not deny that the B.B.C. has in the past shown excellent engineering judgment, not to say foresight.

As might be expected, the bulk of the Report deals with the programme side and also gives detailed accounts of income and expenditure. There is, though, a 10-page section devoted to engineering in which the past year's work is surveyed and plans for the future are set out. In this section there is also a note on present policy in recruitment and training of technical staff. In addition to recruiting ready-trained men, the Corporation is taking in probationary technical assistants. No specific qualifications are expected of these, but they must pass a qualifying examination after an initial 12-week training course.

Colour television is accorded several mentions and it is interesting to read that the main effort of the research department was concentrated on this subject during the latter part of the year. Then there is the rather surprising statement that "the B.B.C. does not expect to introduce regular transmissions in colour within the next two years." This statement, with its inevitable implication that a regular colour service will start shortly after that period has expired, seems rather over-optimistic. In this matter, the Board of Governors appear to be slightly at variance with Sir Ian Jacob, the Director-General, who, in the current number of *The B.B.C. Quarterly,* was (rightly, we think) much more cautious and non-commital. "Nothing," he said, "could be more foolish than a precipitate start with coloured programmes."

* Cmd. 9269 H.M.S.O., 4s. 6d.

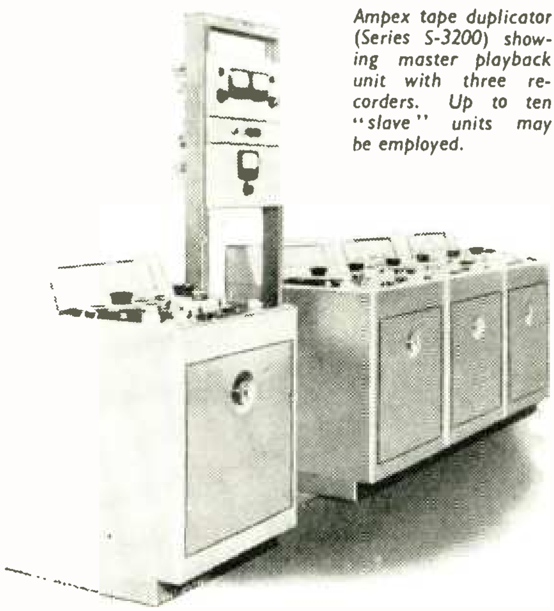
Quality on V.H.F.

AS most of our readers know, the B.B.C.'s plan for reinforcing the present sound broadcasting service with v.h.f. three-programme stations is due to start with Wrotham next May. The remaining eight stations of the first stage of the scheme should be finished within less than two years.

This scheme is described in the Report discussed in the preceding paragraphs, and was also the subject of a talk given recently to the Radio Industries Club by Harold Bishop, B.B.C. Director of Engineering Services. Some disappointment was felt, even among the more realistic, at Mr. Bishop's summary dismissal as "nonsense" of the idea that f.m. broadcasting would automatically bring about a great improvement in quality of reproduction. No doubt, as he said, the idea of giving a bandwidth of 15 kc/s is impracticable for a nation-wide service depending on a long and complicated network of landlines for linking the stations. However, the quieter background and freedom from interference that many f.m. listeners will enjoy will inevitably focus attention on quality of reproduction. It will be a great pity if the B.B.C. does not do all it can, within the bounds of good engineering, to ensure the best possible quality from the new service.

Tape Duplication System

EQUIPMENT for the duplication of recorded magnetic tapes on a commercial basis is now being sold in America. A master playback unit feeding a master amplifier and bias oscillator supplies from one to 10 "slave" recorders. Both tracks of the twin-track



Ampex tape duplicator (Series S-3200) showing master playback unit with three recorders. Up to ten "slave" units may be employed.

master tape are read and re-recorded simultaneously and the tape speed for duplication may be as high as 60 in/sec. This calls for pick-up and recording heads capable of handling frequencies up to 120 kc/s if the original tape carries 15 kc/s at $7\frac{1}{2}$ in/sec. The heads used are designed for supersonic recording and are similar to those which have been developed by Ampex for recording data in flight testing.

No time is wasted in rewinding the master since the machine duplicates in both directions. It has been calculated that, with the full complement of 10 recorders, the output of $3\frac{3}{4}$-in/sec tape duplicates is increased by a factor of as much as 320 over a pair of standard recorders running at the original speed of the recording. (Speed $\times 16$, twin tracks $\times 2$, recorders $\times 10$.)

Total cost for a 10-unit installation is said to be less than $20,000. The makers are Ampex Corporation, 934, Charter St., Redwood City, California.

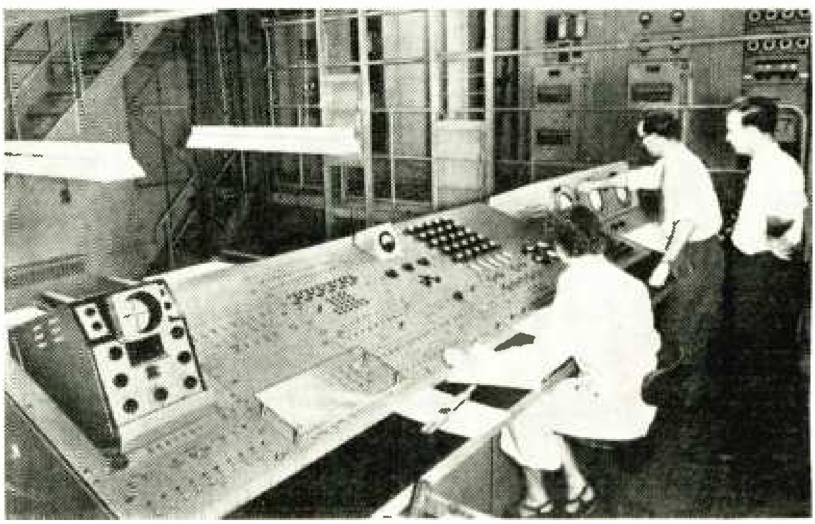
N.B.S. Laboratories

THE new research centre of the American National Bureau of Standards at Boulder, Colorado, was officially opened on September 11th. A series of scientific meetings was held coincident with the opening and included a symposium on propagation, standards and problems of the ionosphere. Among those present at the opening was H. J. Finden, chief engineer of the Electronic Instruments Laboratory of the Plessey Company, to whom we are indebted for the following report.

The Boulder Laboratories, which supplement the N.B.S. facilities in Washington, D.C., include the Bureau's Central Radio Propagation Laboratory transferred from Washington. The C.R.P.L. is the chief American research centre for the study of the troposphere and ionosphere as media for the propagation of radio waves. It also develops and maintains the national primary standards for the complete radio-frequency spectrum. As the nation's central agency for collecting radio propagation data, the C.R.P.L

GUIDED-MISSILE SIMULATOR

The control desk of "Tridac," a large-scale analogue computor built by Elliott Brothers and recently put into operation at the Royal Aircraft Establishment, Farnborough. It calculates the flight behaviour of new types of guided missiles in three dimensions and gives the distance by which they miss the target. Parts of actual missile control systems can be included in the computor, as it behaves as a model of the missile flight and operates in "real" time. The c.r.-tube displays on the right give a representation of the missile approaching its target in three separate dimensions. Computing elements in the machine include large numbers of d.c. amplifiers, and some idea of the techniques used can be gathered from "Electronic Analogue Computing" in our March, 1954, issue.



analyses and disseminates information which is of vital importance for the maintenance of radio services in aviation, shipping and communications.

The Laboratory's studies of frequency allocation and interference affect the establishment and operation of all American broadcasting stations. Data on v.h.f. radio propagation and the development of micro-wave techniques are important to the weather bureau and military aerologists for use in the measurement of upper air temperature, humidity and wind.

Boulder is well situated in that the plains extending hundreds of miles eastwards from the mountains permit many phases in radio propagation research. Use is also made of the nearby mountains and several transmitters are located in the Cheyenne, near Colorado Springs. Frequency utilization research carried out by the N.B.S. is providing valuable data on the effects of noise due to the troposphere as well as the sun and terrain, particularly at frequencies above 50 Mc/s.

"A Guide to Amateur Radio"

THIS book was first published by the Radio Society of Great Britain in 1933; a new edition appeared each year on the opening of the National Radio Exhibition until 1937, when it was superseded by *The Amateur Radio Handbook*. The present publication, styled the sixth edition, follows the tradition of the earlier ones, but is far more comprehensive.

Its aim is to provide the newcomer to amateur radio with up-to-date information on present-day practices and to give expert advice on how to obtain an amateur transmitting licence. There are chapters on learning morse, making simple equipment, abbreviations commonly used in amateur transmission and sundry other subjects not easily found elsewhere.

The publisher's address is New Ruskin House, Little Russell Street, London, W.C.1, and the price is 2s 6d (2s 9d by post).

Fleming's Pre-Valve Work

From Watts to Kilowatts

THE name of Sir Ambrose Fleming is inevitably bound up with the invention of the thermionic diode and this month celebrations are being held to mark the jubilee of the first valve patent (see October issue, p. 474). On such an occasion it is interesting to recall that Fleming had been engaged in radio work for some time before making his great invention, and can be considered as one of the pioneers of the spark transmission era.

Probably his most important pre-valve work was on the design of the famous Poldhu transmitting station, by means of which the Atlantic was first spanned by radio in December, 1901. Here he was acting as scientific adviser to the Marconi Company. Up to this time, radio people had been thinking in terms of the conventional induction-coil spark transmitter, but Fleming realized that a different approach would be needed for a station that would have to radiate kilowatts instead of watts.

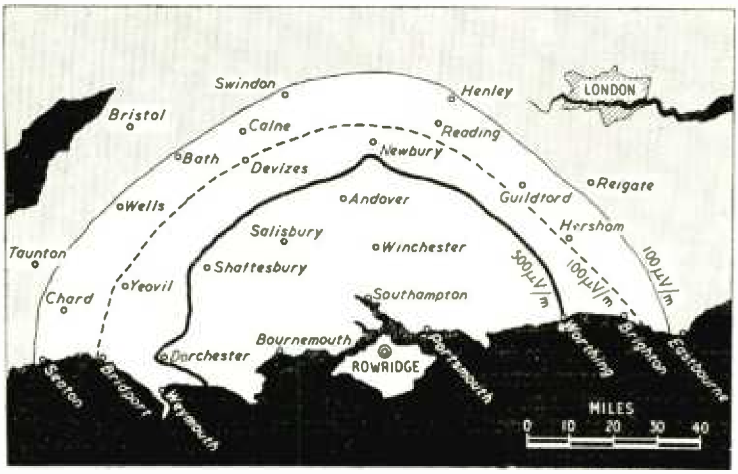
In place of an induction coil he specified an alternator driven by a 25-h.p. engine and working into high tension transformers to give an output voltage of 20kV. This arrangement fed an oscillatory circuit, having the usual spark gap and capacitor, which in turn was coupled by a step-up r.f. transformer to a second similar circuit. Finally came a radio-frequency transformer or "jigger" for coupling to the aerial. Keying the transmitter was achieved by short-circuiting air-core chokes inserted in the leads running from the alternator to the high-tension transformer.

Fleming's contribution to the famous transatlantic experiment is commemorated on the granite column which marks the site of the station on the cliff-top at Poldhu.

SOUTH COAST TV STATION

Coverage of Isle of Wight Transmitter

Provisional field-strength contours of the new Rowridge, Isle of Wight, television station which has been radiating test transmissions for the past fortnight and is scheduled to be brought into service on November 12th. The station, which operates in Channel 3 (56.75 and 53.25 Mc/s), will use initially a temporary mast and aerial system and the service area will, therefore, be restricted. The anticipated 100-μV/m contour for the temporary aerial is shown dotted on this map. This and the other contours are based on a receiving aerial height of 30ft. The temporary station on Truleigh Hill, near Brighton, which has been serving part of this area of the south coast since May last year, will close down.



WORLD OF WIRELESS

C.C.I.R. Interference Investigation • Receiver Exports • Personal and Industrial Notes and News

I.F. for TV

REALIZING that a nationally protected i.f. band for television does not offer protection against interference from other countries, Italy has asked the International Radio Consultative Committee (C.C.I.R.) to study a number of questions relating to interference with television reception and interference to other services by television receivers.

Among the specific subjects to be studied are: the factors which govern the frequencies and amplitudes of undesired receiver responses and the characteristic values for different types of receiver (sound, television, etc.); the methods which can be adopted to reduce these responses without greatly increasing the cost of receivers (the choice of i.f. falls into this category); the best methods for measuring and evaluating local oscillator and i.f radiation; typical values of the amplitudes of these radiations and the variation of their power as a function of distance; the methods whereby a useful reduction in the amplitudes of these radiations could be achieved without appreciable increase in receiver costs and the extent of this reduction.

Information on these subjects is to be submitted by national representative organizations to the C.C.I.R. and, in the meantime, the European Broadcasting Union Document Tech. 3062 (see our July issue) has been submitted as a contribution to the inquiry.

Amateur Radio Show

AT noon on November 24th, the eighth annual Amateur Radio Exhibition, organized by the Radio Society of Great Britain, will be opened at the Royal Hotel, Woburn Place, London, W.C.1, by H. Faulkner, C.M.G., director of the Telecommunication Engineering and Manufacturing Association. The exhibition will continue until November 27th and be open daily from 11.0 a.m. to 9.0 p.m. Admission is 1s.

Members of the R.S.G.B. will be exhibiting a wide range of home-constructed equipment, and the following manufacturers and organizations are taking space: Air Ministry, Amos, Avo, Cosmocord, English Electric, Enthoven Solders, G.E.C., Grundig, Labgear, Magnetic Devices, Minimitter Co., Philpotts, Pye, S.T.C., *Short Wave Magazine*, Taylor Instruments, War Office, *Wireless World*.

Television Exhibition

THE Television Society's annual exhibition will be held from January 6th to 8th in the gymnasium at University College, Gower Street, London, W.C.1. Admission on the opening evening (6.0-9.0) is limited to members, but tickets will be available to non-members for the following two days, when the show will be open from noon to 9.0 p.m. and 10.0 a.m. to 7.0 p.m. respectively. Tickets will be available from the secretary at 164, Shaftesbury Avenue, London, W.C.2.

About 40 exhibitors will be participating. The majority of exhibits will be laboratory equipment provided by manufacturers, but there will be some members' exhibits. The exhibition is concerned more with research and industrial television than with domestic reception.

Middle East Market

TWELVE receiver manufacturers are sharing the B.R.E.M.A. pavilion at the British Trade Fair which is being held in Baghdad from October 25th to November 8th. The firms participating are: Bush, Cossor, Ekco, E.M.I., English Electric, G.E.C., Kolster-Brandes, Mullard, Murphy, Philco, Pilot and Regentone. Other radio and electronic manufacturers participating in the Fair are B.I. Callenders, Ediswan, Ever Ready, Pye, Redifon, Roberts Radio, S.T.C., and Thorn Electric.

Television programmes for the demonstration of receivers at the Fair are broadcast by a temporary transmitter (using a 60-foot mast) installed by Pye.

The Middle East is regarded as one of the most promising markets for British radio equipment. Figures provided by B.R.E.M.A. show that the value of receivers and radio-gramophones exported to this area in the first eight months of this year almost equalled those for the whole of 1953—£364,906, compared with £412,359.

"No Half-way House"

"NOW that colour has become the obvious next move in the development of television, let us learn from our past mistakes and not again hitch ourselves to a wrong system [as in 1946 when we re-adopted 405 lines]. . . . Let us make our own mistakes if we must, but not a mistake just because the Americans have already made it. . . . The system we start with is the system we shall end with, and in my opinion it rules out any public experiment of a compatible or semi-compatible system in Band III with the hope that later we will be able to move colour television to Bands IV and V." This extract is taken from a speech by C. O. Stanley printed in the annual report of Pye, Ltd., of which he is chairman. He concluded by saying that the radio industry, as a whole, could and should provide the answer to the colour TV question, and not wait for any committe, Government or otherwise, to produce a recommendation.

Servicing Exams

OF the 367 candidates who sat for the Radio Servicing Certificate Examination last May, 143 passed, 96 were put back in the practical test and 128 (35 per cent) failed. The percentage of failures in the television servicing exam was lower—28 per cent. Of the 104 entries 55 passed, 20 were put back in the practical test and 29 failed. The examinations, which are conducted jointly by the Radio Trade's Examination Board and the City and Guilds of London Institute, were held at 24 centres for radio servicing and 7 centres for television servicing.

As mentioned in our last issue the 1955 radio servicing exams will be held on May 3rd and 5th (written) and 14th (practical). Those for television servicing will be on May 9th and 11th (written) and June 18th (practical). Entries for the television servicing exam. (fee 3 gns), must be sent to the R.T.E.B., 9, Bedford Square, London, W.C.1, by January 15th, and those for the radio servicing exam. (fee £2 12s 6d), by February 1st.

PERSONALITIES

Dr. W. H. Penley, B.Eng., Ph.D., A.M.I.E.E., who in 1940 joined the Telecommunications Research Establishment and for the past four years has been superintendent, air defence radar, has been appointed senior superintendent, guided weapons, at the Radar Research Establishment, Malvern, which now incorporates T.R.E.

Group Captain R. C. Richmond has been appointed to the London Office of Marconi's Wireless Telegraph Company (Marconi House, Strand, W.C.2) where he will be concerned with the Company's aeronautical radio business. He joined the R.A.F. in 1929, and took the specialist signals course. During the war he was Chief Signals Officer of various commands and was at one time Chief Signals Officer, Air Defence of Great Britain. Prior to retiring, Group Captain Richmond was Commanding Officer of No. 2 Radio School, Yatesbury.

Grp. Capt. R. C. RICHMOND

S. E. ALLCHURCH

S. E. Allchurch, O.B.E., secretary of the British Radio Equipment Manufacturers' Association since 1946, is at the British Trade Fair in Baghdad to take charge of the composite exhibit of domestic receiving equipment by 12 member-firms. During the war Mr. Allchurch joined a new department of the Ministry of Aircraft Production which dealt with the co-ordination of research, development, production and installation of communication and radar equipment for the R.A.F. He was assistant director when he left to join B.R.E.M.A.

William H. Date, B.Sc.(Eng.), M.I.E.E., has retired from the position of head of the Electrical Engineering Department of the Polytechnic, Regent Street, London, which he has held for seven years. He joined the full-time staff of the department in 1913. During the first world war he was a technical officer (wireless) in the Royal Flying Corps and during the last war was lent to the War Office and was attached to the department concerned with the organization of training schemes for Service men at technical colleges throughout the country.

The new head of the Electrical Engineering Department of the Polytechnic is **Dr. D. O. Bishop,** Ph.D., who received his academic training at the Portsmouth Municipal College and then went to B.T-H. at Rugby. After war service as an education officer in the R.A.F. he returned to the Portsmouth Municipal College and became senior lecturer in 1947. He joined the Regent Street Polytechnic as senior lecturer in the Electrical Engineering Department in 1948.

G. E. Middleton, M.A., the new chairman of the I.E.E. Cambridge Radio Group, went to B.T-H., Rugby, after graduating from Cambridge in 1927. During his 21 years with B.T-H., one of which was spent in the U.S.A. under the company's Fellowship scheme, Mr. Middleton was engaged on the design of small motors. In 1948 he went to Cambridge as university lecturer in engineering—his present position.

Major P. L. Barker, B.Sc., this year's chairman of the Northern Ireland Centre of the I.E.E., has been chief engineer of the N. Ireland Region of the Post Office since 1946. After graduating from Birmingham University in 1923 he entered the Post Office Engineering Department. From 1925 to 1935 he was at the Dollis Hill Research Station working on short-wave propagation and from 1936 until he was commissioned in R.E.M.E. in 1940 he was at the Wembley Laboratories.

Dr. J. H. Mitchell, Ph.D., B.Sc., the new chairman of the East Midland Centre of the I.E.E., has been head of research with Ericsson Telephones, Ltd., Nottingham, since 1947. He studied at Bristol University after which he joined B.T-H. at Rugby as a research engineer. During the war Dr. Mitchell was a member of the Government Scientific Research Pool and undertook research on radar and radio navigational aids and v.h.f. communication.

David H. Thomas, M.Sc.Tech., chairman of the North-Eastern Radio and Measurements Group of the I.E.E. for this session, was for eight years lecturer in telecommunications at the University of Nottingham and is now head of the Electrical Engineering Department of the Rutherford College of Technology, Newcastle-upon-Tyne. Before entering the scholastic field he was a research engineer with Metropolitan Vickers whom he joined as an apprentice.

M. I. Forsyth-Grant, A.M.I.E.E., has resigned from International Aeradio, Ltd., with whom he had been chief engineer since 1952, and has joined the board of Racal Engineering, Ltd. Before joining I.A.L. in 1947 he was with E.M.I. Engineering Development, Ltd. **D. W. Morrell,** B.Sc.(Eng.), A.M.I.E.E., who has been sales manager of Racal for the past 18 months, has also been appointed a director.

A. R. Lash, A.M.I.E.E., recently appointed manager-engineer of the Ongar (Essex) radio station of the Post Office, was previously on the staff of Marconi's W.T. Co. and Cable & Wireless. His duties as a radio communications engineer have taken him as far north as Spitsbergen and as far south as the Falkland Islands.

OUR AUTHORS

W. R. Cass, who, with R. M. Hadfield, contributes the article on "Dip-soldered Chassis Production" in this issue, is in charge of the Methods Development Department of Pye, Ltd., Cambridge. Joining Pye from Telephone Rentals, Ltd., in 1951, he was initially connected with power control and television receiver circuitry development. Mr. Cass obtained the Higher National Diploma in Telecommunications at the Regent Street Polytechnic and was an electronics instructor in R.E.M.E. during the war **R. M. Hadfield** studied at Reading University after serving as a pilot in the R.A.F. In 1952 he joined the Methods Development Department of Pye, Ltd., and is now responsible for operational research and work study development.

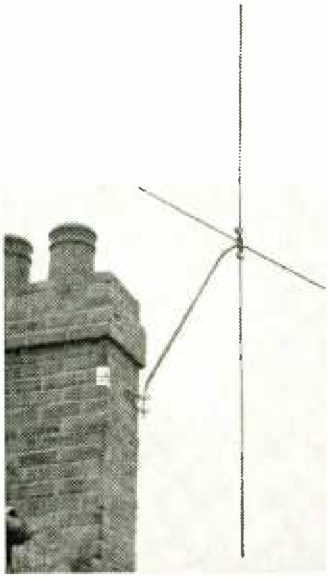
Arieh F. Fischmann, author of the article on the design of a tape recording amplifier on page 564, is at present studying at the Polytechnic Institute of Brooklyn, U.S.A. After training at the Deutsche Technische Hochschule, Prague, he emigrated to Palestine in 1938 where he worked on the design of audio amplifiers. From 1948 until going to Brooklyn he was in the Scientific Department of the Israeli Ministry of Defence on development work in the field of pulse techniques.

H. H. Ogilvy, contributor of the article on the measurement of phase and amplitude in this issue, has been employed in the fire-control section at the Admiralty Engineering Laboratory, West Drayton, Middlesex, since 1951. He is concerned with the development of electronic equipment for fire-control systems and the design of equipment for analysing the performance of servo-mechanisms. After war service as h.f. direction-finding officer in the Navy he took a full-time engineering course at Loughborough College from 1946 to 1950.

IN BRIEF

Broadcast Receiving Licences current in the United Kingdom at the end of September totalled 13,527,864, including 3,677,796 for television and 245,836 for sets fitted in cars. During the month television licences increased by the unprecedented figure of 144,098.

Electronics in Action is the theme of an exhibition of electronic aids to production, design and research being organized by the Scientific Instrument Manufacturers' Association for November 23rd to 25th at the Chamber of Commerce Hall, New Street, Birmingham. Admission to the exhibition is by complimentary ticket, obtainable from S.I.M.A., 20, Queen Anne Street, London, W.1, or the Chamber of Commerce. The exhibition, at which there will be some 20 exhibitors, opens at 2.0 on the first day and 10.0 on subsequent days and closes daily at 8.0.



This combination aerial for television and f.m. sound, exhibited by Belling-Lee at the Radio Exhibition, was by mischance overlooked when our review of the show was compiled for last month's issue

The proposed dates for the second post-war **Northern Radio Show** which the Radio Industry Council plans to hold in the City Hall, Manchester, have been amended: they are now May 4th to 14th, 1955

I.T.A.—The temporary headquarters of the Independent Television Authority, of which Sir Robert Fraser is director-general, are at 12-16, Wood's Mews, Park Lane, London, W.1. (Tel.: Mayfair 6272). The I.T.A. has, so far, been operating from the Post Office headquarters.

The P.M.G. has relaxed one of the regulations included in the recently introduced **Amateur (Television) Licence.** Holders may now transmit messages by telephony or morse without a separate licence provided they are concerned with the technical matter of the visual transmission.

Extending French TV.—The fourth French television station—at Marseilles—was brought into service in September. Operating on 186.55 Mc/s vision and 175.4 Mc/s sound, the 819-line transmitter has an e.r.p. of 50 kW. Initially the station is relying on filmed programmes flown from Paris. On October 15th the fifth television station was brought into service at Lyon-ville which is linked with Paris by radio relay stations. The Lyon-ville 200-watt transmitter operates on 164 Mc/s vision and 175.15 Mc/s sound.

Brit.I.R.E. Awards.—The Clerk Maxwell premium (20 guineas) has been awarded by the British Institution of Radio Engineers to Dr. W. Saraga, D. T. Hadley and F. Moss for their paper "An Aerial Analogue Computer"; the Heinrich Hertz premium (20 guineas) is given to B. E. Kingdon for his paper "A Circular Waveguide Magic-Tee and Its Applications to High-Power Microwave Transmission"; Dr. D. A. Bell receives the Louis Sterling premium (15 guineas) for "Economy of Bandwidth in Television"; Dr. Paul Eisler the Marconi Premium (10 guineas) for "Printed Circuits—Some General Principles and Applications of the Foil Technique" and J. A. Youngmark the Dr. Norman Partridge Memorial award (5 guineas) for "Loudspeaker Baffles and Cabinets." The first award of the Sir J. C. Bose Premium (250 rupees—approx. 18 guineas) for the most outstanding paper by an Indian engineer has been made to S. K. Chatterjee for "Microwave Cavity Resonators—Some Perturbation Effects and Their Applications."

Cable and Wireless announce the opening of the first direct **Radio-telephone Service** between Aden and India on October 15th. A radio-telephone service between Aden and London and thence to Europe and north America is also provided by the company via its station at Nairobi.

The **Radar Association**, which, although originally formed as a social link between ex-R.A.F. "radar types," is now open to radar engineers and technicians in the other Services and industry and holds regular lecture meetings. The second meeting of the 1954/55 session will be held at 7.30 on November 10th in Theatre No. 1 of the Lime Grove Television Studios, Shepherds Bush, London, W.12, when Group Capt. Philip Dorté and John Elliot (B.B.C.) will deal with the production of the television film "War in the Air." Tickets are necessary.

E.I.B.A. Ball.—The annual ball in aid of the funds of the Electrical Industries' Benevolent Association will be held at Grosvenor House, Park Lane, London, W.1, on November 12th. Tickets, price 2½ guineas, are obtainable from the association at 32, Old Burlington Street, London, W.1.

Two lectures on the design, construction and erection of **Television Aerials** are to be given by P. Jones, of Aerialite, Ltd., at the Gloucester Hotel, Aberdeen, on November 3rd and 4th at 3.0. Admission tickets are obtainable from Aerialite, Ltd., Castle Works, Stalybridge, Cheshire.

The Electrical Engineering Department of the **Oldham Municipal Technical College,** which is now occupying a new building, offers part-time day or evening courses in radio servicing (3 years), television servicing (2 years), telecommunication engineering and for the Higher National Certificate.

A series of eight lectures intended to present an up-to-date account of **Information Theory** and its implications in the field of communication engineering is to be given at the College of Technology, Manchester, on Friday evenings, beginning January 21st. The fee is 30s. Twelve-lecture courses on "Automatic Control in Industry," which began on October 26th, and "Transient Electrical Phenomena" beginning on January 13th (fees 35s) are also provided by the College.

Among the new courses provided at the recently opened engineering block of the **N.E. Essex Technical College** at Colchester is a year's evening course in electronics and measurements for the H.N.C. Part-time day and evening courses in preparation for the radio service work certificate of the City and Guilds of London Institute are also provided by the college which has a well-equipped radio servicing laboratory.

Information additional to that given in our September issue (p. 439) regarding courses provided by the **N.W. Kent College of Technology** (previously the Dartford Technical College) has been received from the head of the Electrical Engineering Department. An evening course in television servicing is provided and as an alternative to the evening course in radio servicing a part-time day course has now been introduced.

Properties of **Glass Reinforced Plastics** and their applications in various industries are dealt with in the 256-page book "Glass Reinforced Plastics," edited by Phillip Morgan and published for *British Plastics* by our Publishers, price 35s. One of the chapters deals with glass fibre laminates in the electrical field.

The second edition of Kenneth W. Gatland's book "**Development of the Guided Missile,**" which has been completely revised and enlarged, includes an appendix giving details of the telemetering equipment used in British missiles. Published for *Flight* by Iliffe and Sons Ltd., this 292-page book costs 15s.

Philips have arranged for the future distribution in this country of their three journals *Philips Technical Review*, *Philips Research Reports* and *Communication News* to be undertaken by Cleaver-Hume Press, Ltd., 31, Wright's Lane, Kensington, London, W.8.

BUSINESS NOTES

Transmitters for the first three stations of the Independent Television Authority have been ordered from **Marconi's.** The vision transmitters will have a power of 10 kW and the associated sound equipment 2.5 kW. They will, of course, operate in Band III. To expedite the start of transmissions from the London station at Crystal Palace prototypes of the transmitters are being lent to the I.T.A.

International Aeradio, Ltd., of this country, and Adalia, Ltd., of Montreal, Canada, have entered into an agreement to pool their knowledge, experience and staff in order to offer consultancy services to governments and industry throughout the world. Their first joint contract is for the planning of a telecommunications system for the Creole Petroleum Company, of Venezuela.

Claude Lyons, Ltd., the well-known manufacturers of Variac transformers and importers of American laboratory instruments, have acquired a new factory at Valley Works, 4-10, Ware Road, Hoddesdon, Herts. (Tel.: Hoddesdon 3007.) The London office has closed, but the head office and works will remain at 76, Oldhall Street, Liverpool. The new factory houses the research and development staff and provides a repair and recalibration service.

The Service Department of **Baird Television** has been transferred to 308, Battersea Park Road, London, S.W.11 (Tel.: Battersea 7838). All correspondence relating to servicing and replacement parts for television receivers and Baird tape recorders should be sent to this address.

The West Bromwich firm of spring manufacturers, George Salter & Co., Ltd., have built a "**Dry Room**" to Ministry of Supply regulations for the packing of electronic equipment under conditions which ensure that it is not only impervious to outside climatic changes but also that each piece of equipment is thoroughly moisture-free before being packed.

The recently opened factory of **20th Century Electronics** at New Addington, Surrey, has now been extended, doubling the floor space. The manufacturing and research sections of the cathode-ray tube department, formerly at Dunbar Street, London, S.E.27, are now at New Addington, where new plant has been installed for the production of multi-gun tubes. All correspondence should now be addressed to King Henry's Drive, New Addington, Surrey. (Tel.: Springpark 1026.)

Electric Audio Reproducers, Ltd., manufacturers of sound reproducing equipment, of 17, Little St. Leonards, Mortlake, London, S.W.14, have opened a new factory at Worton Road, Isleworth, Middlesex. The development section will occupy part of the factory, but the offices and service department will remain at Mortlake.

A new branch office and depot at 2, St. Nicholas Buildings, Newcastle-upon-Tyne, 1, has been opened by the **Telegraph Construction and Maintenance Company.** The company has also opened a London sales office at Norfolk House, St. James's Square, S.W.1.

Hudson Electronic Devices, Ltd., of Appach Road, London, S.W.2, have appointed Pendry & Kennedy (Electronics) of 6, Coed Celyn Road, Derwen Fawr, Swansea, to handle the land sales and services of their v.h.f. radio-telephone equipment in the south Wales area.

Transvision, Ltd., has been formed by B. J. Martindill, until recently general manager of Wolsey Television, and F. Gould, for the production of television and v.h.f. aerials and accessories. The address is 118, Denmark Hill, London, S.E.5. (Tel.: Brixton 6551.)

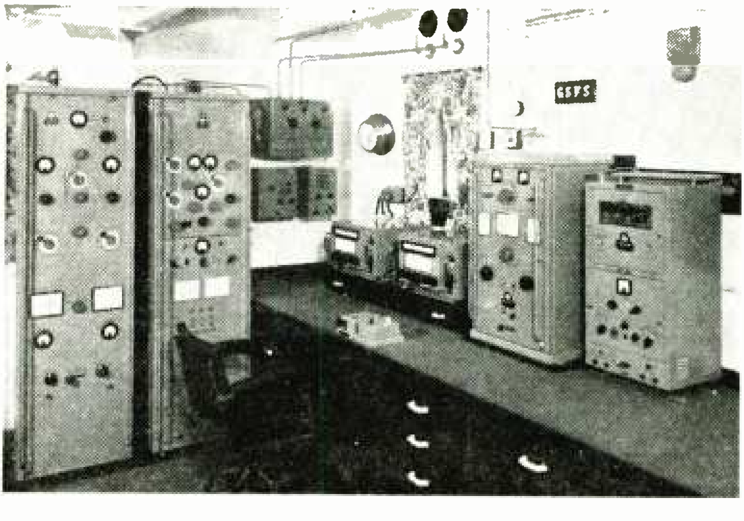
FOREIGN TRADE

Redifon, Ltd., has received a substantial order for marine radio equipment on behalf of the Soviet Fishing Authority. The equipment, comprising transmitters, all-wave receivers, combined medium- and short-wave direction finders and ancillary units, will be installed in twenty deep-sea fishing vessels now under construction for the Soviet Union at Lowestoft, Suffolk.

Haiti.—The British Embassy at Port-au-Prince has received an enquiry from a Government source in Haiti for the supply of robust, cheap battery receivers for use in rural areas. The proposal is to supply about 500 sets for domestic use or for use by small groups, or, alternatively, to equip a smaller number of centres with larger receivers designed for communal use. Manufacturers are invited to send details of their offers direct to H.M. Consul-General, British Embassy, Port-au-Prince, Haiti, W. Indies.

Salvador Agency.—Almacen Liverpool, Calle Ruben Dario-32, San Salvador, are interested in securing an agency for British-made domestic receivers. Transmitters in the western hemisphere operate on medium waves and the mains voltage in Salvador is 110 (50 c/s).

Colombian Agencies for British-made components and accessories are sought by Almacen Radion, Guillermo Ibanez, Calle 36, No. 41-78, Barranquilla, and J. Angers, Apartado Aereo 913, Barranquilla.



WIRELESS ROOM in the 3,300-ton cable ship Recorder, *the latest and fastest vessel in the Cable & Wireless fleet of eight. Marconi's installed the radio-communication equipment and Kelvin & Hughes the radar and echo sounders. An aerial-splitter system permits the use of 35 broadcast receivers in the ship without interference from the transmitters.*

Dip-Soldered Chassis Production

Simplifying the Assembly of Sound and Television Receivers

By W. R. CASS,* H.N.D., Grad.I.E.E., and R. M. HADFIELD,* B.A.

IN the endeavour to reduce the cost of mass-produced electronic assemblies it became apparent that a simpler way of making electrical connections was needed. Conventionally, pliers are used to anchor connecting wires to tags and then solder is applied with an iron to make these mechanical connections into sound electrical joints. In a television receiver there are between 500 and 800 such connections to be made.

After costing several systems it was decided that the most suitable basic technique was one in which all connections are made simultaneously with a single dip in hot solder. For this method it is convenient to bring all points of interconnection into one plane. This is done by laying all the components on an insulated board and inserting their pre-formed wire ends into holes so that all points to be soldered are on the underside. Once such a layout has been adopted it is a short next step to use a printed wiring pattern on the underside of the chassis to replace the wire links which connect one part of the circuit with another.

Superficially it might appear that the addition of printed wiring is of little value, since it obviously costs more than the wires it replaces and only saves a small amount of assembly time. However, this is not so, because the most important function of printed wiring is not, as its name suggests, to eliminate wire links, but to provide a base for dip-soldering. At each component junction a piece of copper foil surrounds the wires, even if there is no circuit line joining this point to another point. This piece of copper picks up solder, which links the adjacent component wires. Besides assisting the soldering of electrical connections, small areas of copper foil left on the board provide points to which the heavier components can be secured in the dip-soldering operation. This saves much mechanical fixing with screws, brackets, etc. A dip-soldering system that does not use a printed wiring base must make use of eyelets at all points where a soldered connection is required. The insertion of these eyelets costs more in material and labour than does a printed wiring board.

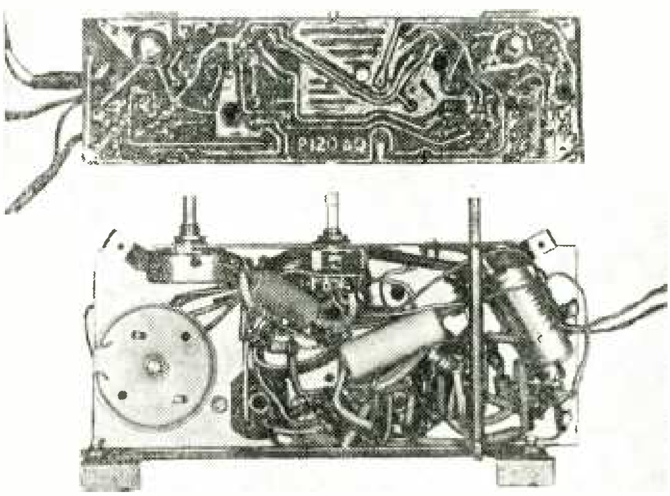
Four times as many sound or television receiver chassis can be made by an assembly line using the dip technique as can be made by one using hand assembly methods. This means not only lower labour costs but reduced overhead costs per unit as well. Savings of the last-mentioned include supervision, factory space, heating, lighting, power and soldering-iron maintenance. It must be emphasized, however, that the assembly cost in mass-produced television and sound receivers is only a small fraction of their total cost. Thus, even the 300 per cent improvement in chassis assembly productivity achieved will only result in overall cost savings of the order of 2-3 per cent.

Material savings result from the elimination of nearly all brackets, tag strips, sleeving, wire, screws, eyelets and rivets. These savings more than offset the extra cost of the copper-faced Bakelite chassis. Some of the savings are possible because the mechanical attachment of many components becomes unnecessary: all forms of screening can be dip-soldered, and trimmers, valveholders, transformers and large capacitors need no additional fixing. Other savings result from the fact that much insulation can be dispensed with when a non-conducting chassis is used. At present this means that neither sleeving nor insulated resistors are necessary in the receivers, and ultimately it will result in cheaper valveholders and transformers as well.

The elimination of the above items plus the use of a laminated Bakelite chassis results in considerable weight saving; this reduces the problem of adequately securing the assembly in a cabinet. The weight saved on the chassis varies from 1lb on a small sound receiver to 4lb on a television set.

Generally, electronic units made by this technique and using conventional components show little reduction in size. It is hoped, though, that with the efforts of the component designers taking advantage of this

* Pye, Ltd.



Underside view of a 4-valve battery receiver made by the "dip-circuit" technique with (below) the same chassis made by conventional methods. The top of the dip-soldered chassis is shown in the picture at the top right of page 538.

new approach to manufacture, drastic reduction in equipment sizes will be achieved.

A description will now be given of the manufacturing methods used to produce a complete "dip-circuit" chassis. First of all the production of the chassis circuit board.

Bakelite faced with copper foil is supplied in 9-in wide strip. This is fed through a piercing and shearing tool which makes some of the larger holes and cuts the material into pieces the size of the chassis blank. Two of the holes made at this stage are for location during printing and during the subsequent major piercing operation.

An acid-resistant ink is next printed on to the copper face of the boards in the pattern of the circuit. The boards are placed, ready for etching, in special acid-proof racks as they leave the printing stage. The printing is done by the silk-screen method. This equipment is cheap and is suitable for work on Bakelite. Output at the rate of 150 circuit boards per hour is maintained with standard equipment, modified to enable the operator to raise the screen frame by a foot pedal. This has reduced fatigue and permits a less complex pattern of hand motions. The operator's hands are not required for lifting the screen frame and so are left free to handle the squeegee and the circuit boards.

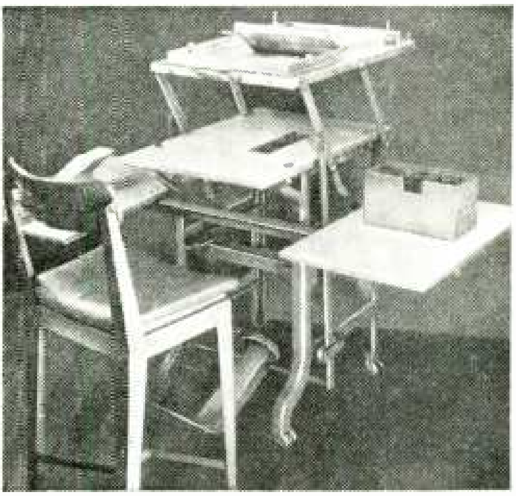
Preparation of the silk-screen master from the photographic plate.

Etching the Circuit

Etching is carried out in tanks containing a solution of nitric acid and copper nitrate. A constant balance is maintained between the acid solution and the weight of copper dissolved in it by calculating the quantity of acid required to etch one chassis and then adding a quantity of acid appropriate to the number of chassis being etched, plus a fixed percentage. The size of this fixed percentage determines the rate at which etching takes place. A 10 per cent excess of acid (which means, in effect, a 10 per cent wastage) gives an etching time of six minutes, using a solution 50 per cent by weight nitric acid in water. This acid wastage varies inversely with the etching time but prolonging the etching time beyond six minutes saves very little acid. Fresh acid is added at the top of the tank and the outflow is taken from the bottom. Turbulence created during etching ensures that the acid is well mixed.

Ferric chloride is sometimes used for etching purposes but we are using nitric acid because it has the following advantages. (a) Four hundred chassis can be etched in six minutes using very simple equipment; no mechanical agitation is needed. To achieve this rate with ferric chloride very expensive equipment is required. (b) Ferric chloride is awkward material to handle and requires heating, stirring and the addition of hydrochloric acid to bring it into solution, whereas nitric acid is readily diluted with the necessary water. Against this must be set the problem of fumes when using acid; this, however, has been solved by using standard extraction equipment. (c) Nitric acid is 30 per cent cheaper to use than ferric chloride.

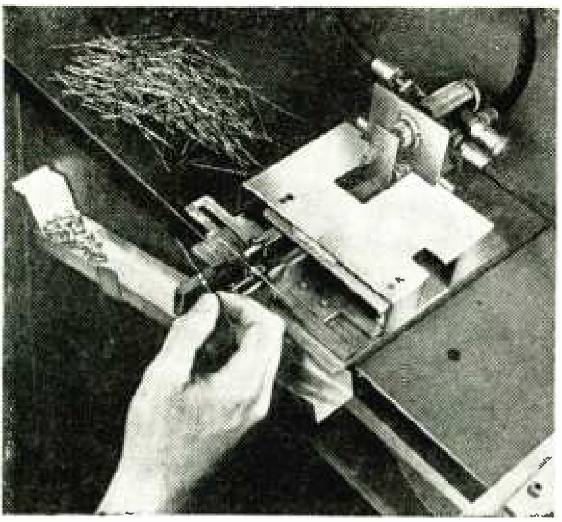
After etching, the boards are rinsed in water, neutralized in a 2 per cent solution of ammonia and then rinsed again. It was decided to use ammonia in preference to sodium bicarbonate because ammonium salts are volatile, and if traces remain on the board they are dispersed during dip-soldering. Next, the circuits are pierced to take the component wires and the valveholder contacts. Finally, the circuits are placed in a trichlorethylene degreaser to remove the resistant ink.

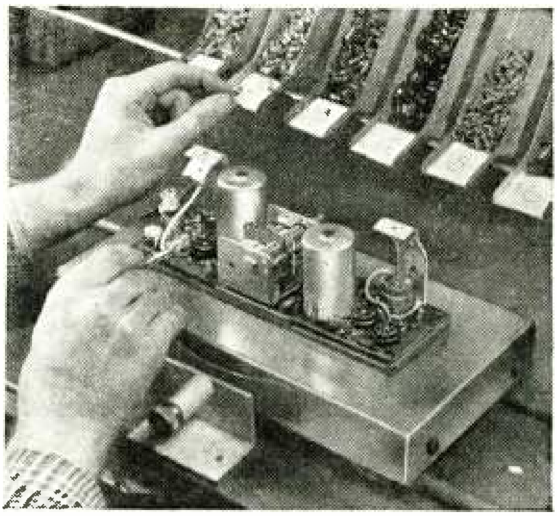


Silk-screen printing machine with foot-operated frame.

To make the maximum use of the labour-saving possibilities of "print-dip" technique, the wire ends of the resistors and capacitors are pre-formed suitably for direct insertion into the holes in the chassis. A machine to do this has been made which cuts and bends both wires of the component and which can be hand fed at rates of 1,200 to 1,500 components per hour.

The machine is driven by a compressed air cylinder linked to an air valve in such a way that it operates automatically once the air is turned on. It can be set to any speed to suit the operator and the position of the cutting and forming blades is continuously adjustable to leave any wire length that is desired. When they have been formed the components drop into a container. This container is one of the interchange-





Left: Wire cutting and forming machine for resistors. Right: Part of the assembly line for the 4-valve dip-soldered battery set whose underside is shown in the picture on page 536.

able component storage bins used on the assembly line.

Both the rigid layout imposed by the system and the mechanical pre-forming of the components have tended to standardize assembly operations. This improves assembly efficiency and leaves less to chance in positioning the components—it is virtually impossible to build a short circuit into a printed assembly.

Assembly Trolleys

The fact that only one side of the chassis has components on it has simplified assembly cradles and work on the chassis. Material layout has also been helped because the pre-formed components are more compact and require less bench space. The assembly cradle is a rubber wheeled trolley free to run in a channel on an ordinary wood-topped bench. A shallow rectangular frame is used to locate the chassis on top of the trolley. At each work station there is a spring loaded catch for retaining the trolley while it is being worked on. Trolleys from the end of the line are returned to the beginning on a sloping shelf at the back of the assembly bench.

Component storage bins are of the gravity-delivery type with a protruding lip at the bottom to facilitate the selection of material. The lip of the bin is clear of the bench so that, with the thumb under the lip and the forefinger on top, the resistor or capacitor can be drawn to the edge and held between the thumb and finger as it comes clear of the bin.

For dip-soldering a machine has been devised to do the job because, while it can be done quite simply by hand, the quality of the soldered joints is so important and the factors which can vary are so numerous that to obtain consistent results a mechanical method is necessary.

To use the machine the operator simply hooks the chassis on to a bracket and presses a button. The machine then moves the chassis to a flux spray position, then gives it two dips in the solder. During these dips, and for a short period after each one, the chassis is vibrated. This vibration helps to break down any oxide film on the circuit or on the components and also removes any excess of solder.

Finally, the machine ejects the completed chassis and returns it to the operator for inspection.

There are some benefits accruing from the introduction of "dip-circuits" whose value it is difficult to assess, but which may turn out to be important in the long run:—

1. Storage and material handling will be less for the smaller, non-insulated pre-formed components.

2. Training time will be cut down, as workers will no longer require skill in the use of pliers and soldering irons.

3. Testing, inspection and repair time will all be cut because of the reduction in wiring errors and the ease with which such errors that remain can be detected on the accessible layout that "dip-circuits" provide.

4. The quality of the finished product will be improved and more easily controlled. The rigid layout of the printed wiring assembly will impose a uniformity not obtainable with conventional assembly methods. This uniformity will, of course, mean better quality as closer adherence to the designed performance will be achieved. It must also be emphasized that the mechanization of the soldering operations will greatly improve the reliability of the units.

5. The servicing of a "dip-circuit" chassis will differ only slightly from that of a normal one. Faulty components can be cut away from the top side of the chassis, or, alternatively, the components can be removed by heating the soldered joints on the reverse side with a small-wattage iron and pulling out the components. As there is no mechanical wrapping of the component wires and no wiring mistakes are possible, servicing time will be reduced. In cases where, due to misuse, the printed wiring pattern has been broken, a wire replacement can easily be inserted or the gap can be closed by the use of a soldering iron.

There will be other changes that will be less welcome. For example, in the immediate future purchasing policy will be less flexible because the rigid circuit layout will not permit any change in the middle of a production run to components of a very different shape or size. Later on this may bring benefits by stimulating standardization among the products of different manufacturers.

Another consequence of adopting the printed layout will be an increase in the cost of circuit modifications after tooling. Alterations to the circuit pattern are quite a simple matter, but modifications to the chassis piercing tools are expensive. This does not mean that there need be serious delay in the introduction of modifications, because temporary expedients, such as the use of a drill jig, are available. However, it does mean that adequate pre-production planning will be at an even greater premium than it is at present.

Until a new process has been widely adopted there is usually a period when many firms are developing their own methods more or less in secret. This has certainly been the case with printed circuits. As a result, there has been no agreement yet in this country on standards of quality for the materials used, or on the dimensions and style of components for "dip-circuit" application. Unless the electronics industry, through its consultative bodies, soon agrees on the general direction that component development is to follow, the cost and quality of components is likely to suffer.

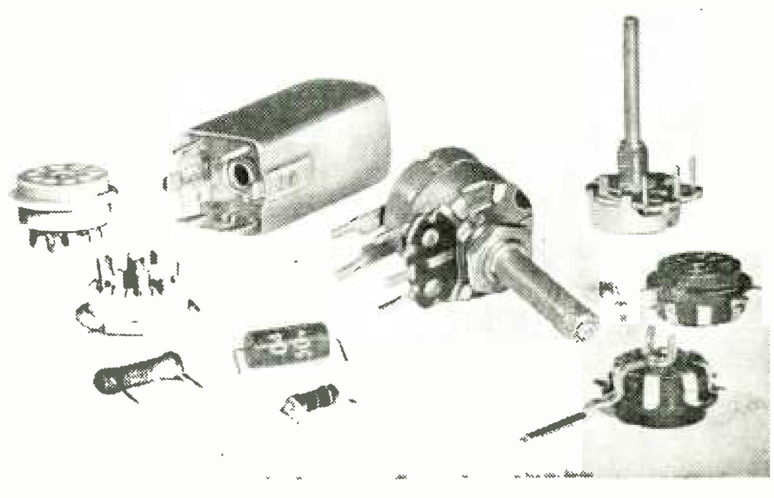
Automatic Component Insertion

With the advent of dip-soldering and printed wiring three-quarters of the possible saving on existing assembly times will have been achieved. If machinery were developed for the automatic insertion of components these times could be reduced by about a further one-tenth. This machinery would be highly specialized and would cost more to develop and install than would the equipment necessary for producing "dip-circuits." In view of this it is not likely that such equipment will repay development in this country, except in cases where production runs are very large or where part of an automatic component manufacturing process can be combined with the automatic insertion process.

A further argument against developing specialized machinery solely for component insertion is that the biggest scope for reducing the cost of electronic equipment lies with the component manufacturers. This may result in the development of components very different in shape and size from those commonly used at present. One such development, called "modular assembly" has already been tried in the U.S.A.* In this system small capacitors and resistors are printed on ceramic wafers and the wafers are assembled in tiers supported by wires; the wires also provide electrical interconnection. This tiered assembly or "module" is surmounted by a valveholder. Several such assemblies go to make a complete circuit and they can be assembled in one unit on a printed wiring base. In the opinion of the authors it has not yet been proved whether such an arrangement is really economical or has a very wide application, but it indicates the possibilities for a radically new approach to the problem of manufacturing electronic components.

Finally, there is the possibility of reducing the costs of component manufacture. In their essentials electronic units consist of four types of elements: capacitors, resistors, inductors and valves. With the possibility of transistors replacing valves there are left for consideration the first three types of elements. Reducing these in turn to their essentials, their costs are made up of: manufacturing labour, conductive and resistive material, some form of casing and terminals for assembly. Of these, the first offers scope for economy but the last two, performing no essential electronic function, are the main targets for material cost reduction. The largest labour savings will result from cheaper methods of producing inductors. In this field printed circuits will have an important part to play. The smaller inductors can be incorporated in the circuit pattern or printed on a separate circuit board which plugs into the base board. This method has been used commercially in the U.S.A. for the coils of a television i.f. strip. This strip was marketed as a separate unit, comprising three valves, five coils and the associated resistors and capacitors assembled on a printed wiring base. Larger inductors such as chokes, transformers and deflection coils need to be printed on a thinner, flexible base material and folded to make them into working units.

The elimination of non-essential material on resistors and capacitors can be achieved by the use of basic components without wires and with simpler insulation. To this end it is sometimes possible to include several electronic elements in one package. The printed circuit "couplet," which consists of a thin, flat, ceramic plate with multiple resistor and capacitor patterns silk-screen printed on one side, has done this and so also has "modular assembly." Multiple capacitor packs and transformers with several voltage tappings use this principle to a lesser extent. However, this trend towards multiple components conflicts with the standardization essential to economic mass production; it also increases the cost of rejects during manufacture. A more flexible system, making use of the physical configuration of either the "couplet" or the "module" but having standard or new-type basic components inserted during the final production process, will perhaps prove to be the best answer.



Components with electrical terminations made suitable for dip-soldering use.

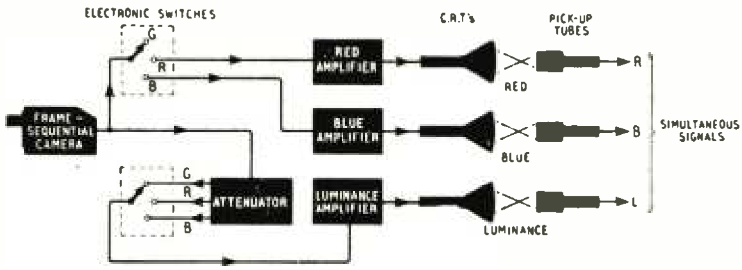
* See *Wireless World*, April, 1954, p. 185

Colour Camera Convertor

Adapting Frame-Sequential Pictures to Simultaneous Transmission System

THE frame-sequential system of colour television has been out of favour lately as a possible means of establishing a public colour service because of its non-compatibility—not to mention the difficulty of the rotating discs at the receivers. Despite this it is known to give very good colour pictures. Moreover, although the system as a whole may be unsuitable, there is a particular part of its equipment—the colour camera—which can offer some very definite advantages if incorporated in a simultaneous compatible system such as the one now operating in America.* This camera is much smaller and lighter than the three-tube type normally used and is simpler and less costly to produce. Having only one pick-up tube, it avoids the necessity for matching and registering the three separate tubes and also the need for three separate amplifiers. Moreover, it avoids the complex optical system which makes turret changes difficult and causes loss of light in the three-tube camera.

To permit the frame-sequential camera to be used in a simultaneous system a device called the "Chromacoder" has been produced. This operates on the three colour-component signals from the camera—red, blue and green—in such a way that they appear not in sequence but simultaneously. The original Chromacoder" was designed in America by the Columbia Broadcasting System and the General Electric Company, but recently a new version of the device has been demonstrated in this country by Emitron Television, a subsidiary of Electric and Musical Industries.



The principle of the Emitron convertor (see block diagram) is to take the sequential red, blue and green signals from the camera and pass them to an electronic switching system which distributes the red signals to one c.r. tube, the blue signals to another and certain proportions of all three signals to a third tube. These three c.r. tubes are then viewed by three pick-up tubes whose outputs give the simultaneous signals. Although the effect on the screens of the c.r. tubes is still sequential, the pick-up tubes store the images on their mosaics until they are scanned off and consequently the outputs become truly simultaneous.

The convertor is arranged in this particular way to make it suitable for the N.T.S.C. type of transmission system, which sends out a luminance signal to provide a monochrome picture for existing black-and-white receivers and two colour-difference signals to provide colour information. In the convertor the luminance information (a mixture of all three colour components) is received in sequential form by the appropriate c.r. tube and the associated pick-up tube integrates it into a complete luminance signal simultaneous with the colour signals. (The green component is recovered later at the receiver by subtracting the red and blue signals from the luminance signal.)

One advantage of this scheme over the three-tube type of "simultaneous" camera is that the three colour components which are added to form the luminance signal all come from the same pick-up tube—the camera tube. They are therefore registered perfectly with each other and no loss of definition occurs through mis-registration. In the three-tube camera, however, the three components come from separate tubes and this difficulty of registration has to be overcome. It is, of course, particularly important for the luminance signal to have good definition because the human eye has great acuity for fine detail in the form of brightness changes, and it is the luminance information which really controls the sharpness of the final picture. On the other hand, the eye is not very sensitive to detail in colour, so that as the red and blue signals from the first two pick-up tubes only provide colour information, not luminance, there is no need for these two tubes to be so accurately registered and give such good definition.

At the demonstration the frame-sequential camera was operating on 405 lines, interlaced 2:1, with 150 frames per second, and the bandwidth was 9 Mc/s. The three pick-up tubes, however, which were C.P.S. Emitrons, were scanned with 625 lines, interlaced 2:1, at 50 frames per second. No doubt the difference of standards helped to avoid the line-beating patterns which might have been caused by the interaction of two similar rasters, but E.M.I. say that there is no reason why two identical standards should not be used.

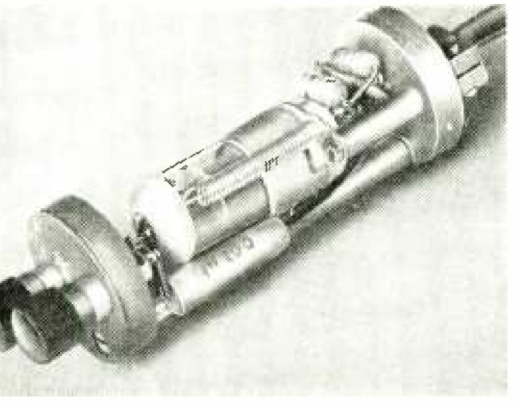
One inherent drawback of the convertor is that the storage in the pick-up tubes is liable to cause blurring of quickly-moving objects in the picture. This was particularly noticeable at the demonstration when the camera was panning from one subject to another. The colour rendering, however, was very pleasing on all the display systems used, which included an R.C.A. tri-colour tube and a three-tube projection unit.

* See *Wireless World*, November, 1953, p. 524.

Cathode-Follower Probe

For Test and Measurement

By SYDNEY H. FINN



Construction of the probe with the Paxolin-tube cover removed.

IT is usually required of an electronic measuring device that its input impedance be high, but the value to which an amplifying-valve grid resistor may be raised is limited, particularly so if potentiometer attenuation is used. Some improvement on a continuously variable input potentiometer is possible by fitting a switch and equalizing each position independently. If a high value of resistance is used then a screening cover will probably be required to reduce hum pick-up. For some purposes the clicking of a switch would be a disadvantage.

In quite a number of instances it is possible, by fitting a probe housing a cathode-follower valve, to more or less completely solve the problem. The attenuator in the main instrument may then be of some conveniently low value, whilst the input impedance to the probe will be very high. By using a probe extremely short leads to the measuring point are possible, although the cable-form to the main instrument can be relatively long.

One such practical arrangement is shown in the circuit diagram and picture. As for some purposes it may be required to use a valve with a high cathode current the load resistor R_L is fitted to the socket on the main instrument, thereby avoiding undue heating of the probe. It also allows some flexibility, in that the same value of load resistance may not be desired for every instrument to which the probe is attached. The six-way socket is so wired that this resistor does not shunt the input circuit after removal of the probe. The cathode follower anode, being at earth potential (to a.c.), acts as a screen around the grid and there is normally no need to use a screened probe. In any case, the outside of the probe should be insulated, otherwise it may be a nuisance when working inside apparatus (particularly if using very short connecting leads) and one wishes to lay the probe down inside the chassis.

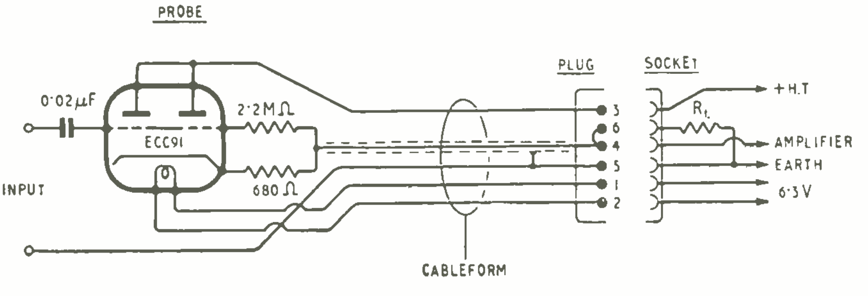
The mechanical arrangement is not necessarily the best that could be devised, but is very simple if a lathe is available. The base is made of an aluminium alloy, the outer sleeve of Paxolin tube, and the terminal plate of loaded ebonite (Caramot RM70). A cable clamp is an essential if the probe is to be much used and would ideally be formed as part of the base. Actually it was modified from an already-existing device. A very simple type of clamp can be made using a short piece of angle with one end flanged. This is screwed to the base (by means of the flanges) so that the cable lies inside the included angle. Twine is then bound tightly round the whole, preferably within grooves filed on the edges of the angle. These grooves prevent any subsequent movement of the binding.

Choice of Valve

The particular valve fitted was used only because it was to hand and was adequate for the job. A triode-connected pentode of the Z77-EF91 type would do just as well. If for some reason a valve with a higher cathode current is required, the Osram N78 has a B7G base and a maximum anode dissipation (triode connected) of 12 watts. As screening will, in general, be unnecessary, ventilation can be provided without compromising the design electrically. The valve retainer is a Carr Fastener type 77/264.

Instruments designed to have an input impedance high in relation to, say, 600-Ω lines, are commonly found in the laboratory. This input impedance, which may be around 100kΩ, will often be the limiting factor in an otherwise first-class instrument. The fitting of a probe may greatly increase its scope, and perhaps even save the cost of a new instrument, if for any reason higher impedance networks must be dealt with. The author has found this probe a very useful instrument to use in conjunction with a c.r. oscilloscope.

If absolute measurements are required an initial calibration will have to be carried out, as the voltage output is somewhat less than the input.



Circuit of the probe with plug and socket connections. The load resistor R_L is 22kΩ nominal. For the plug and socket, Painton types 500693 and 500680 were used.

Miniature Bedside

Two-Valve A.C. Mains Circuit With Pre-set Tuning

THIS article describes a receiver recently built by the author for bedside use by a child. It is a very simple a.c. mains-operated receiver employing two valves and giving approximately 1 watt output from a 5-inch diameter loudspeaker. Tuning is pre-set, the local Home Service or Light Programme being selected by a 2-way switch. The receiver is relatively inexpensive to build and all the components, including the tuning coil, are standard commercial products which are readily obtainable. Where local-station reception is all that is required and great volume is unnecessary, the receiver is suitable for general domestic use.

As the receiver was intended for a child, consideration was first given to the use of batteries, but battery-replacement cost can be serious when a receiver is used in this way and mains operation was decided in spite of the additional bulk and first cost of mains transformer, rectifier and smoothing components. An earthed chassis was considered essential and a mains h.t. transformer is used although receivers with a filament transformer or an l.t. dropping resistor are smaller and cheaper. As the receiver is mains-operated it does not greatly matter if it is accidentally left on, but if this occurs the indicator lamp at the front serves as a reminder.

Great volume is not required from such a receiver and 1 watt output is quite adequate. This can be obtained from a miniature valve of the 6AM6 type for approximately 1 volt input, and two valves of this type give all the gain necessary for local-station reception. A second 6AM6 is therefore used as a leaky-grid detector, the two valves being coupled by a "starvation" circuit as shown in Fig. 1. This particular form of coupling has been described elsewhere[1,2] and is adopted because it is economical of components and gives high gain. It is not desirable, however, to carry the process of starvation too far, otherwise the receiver does not function very well on strong signals. V1 is a leaky-grid detector and, on receipt of signals, generates a negative bias on the control grid approximately equal to the carrier amplitude. If the valve has a very low screen-grid potential, the bias produced by a strong signal may cause the valve to operate on a markedly non-linear part of the I_a-V_g characteristic producing unpleasant harmonic distortion. To minimize this effect the screen-grid potential must be kept high to give V1 an adequate grid base; this sets an upper limit on the value of anode load which can be used and prevents full exploitation of the starvation circuit. In the compromise solution adopted by the author, the screen grid is operated at 40 volts and results are satisfactory provided that the input signals are reasonably small. If a long outdoor aerial is used, or in regions of particularly high field strength, it may be desirable to reduce the input to the detector. A convenient way of doing this is described later.

The required screen-grid potential of 40 volts is obtained in the following way. This potential is also that of the cathode of V2 and, for optimum results, V2 should consume approximately 12 mA. This gives the value of R_7 as $40/(12 \times 10^{-3}) = 3.3\,\text{k}\Omega$. The value of R_3 is now chosen to give a screen-grid potential of 40 volts. The value used by the author was 820 kΩ but others may find a slightly different value is required, dependent on the characteristics of the particular valve used as detector.

Gain Control

One of the difficulties of a circuit such as that shown in Fig. 1 is that of controlling gain. It is not possible to use a potentiometer in the coupling between the valves without upsetting operation of the starvation

[1] W. K. Volkers, "Direct-coupled Amplifier Starvation Circuits," *Electronics*, March, 1951.
[2] H. E. Styles, "Sensitive Two-valve Receiver," *Wireless World*, May, 1953.

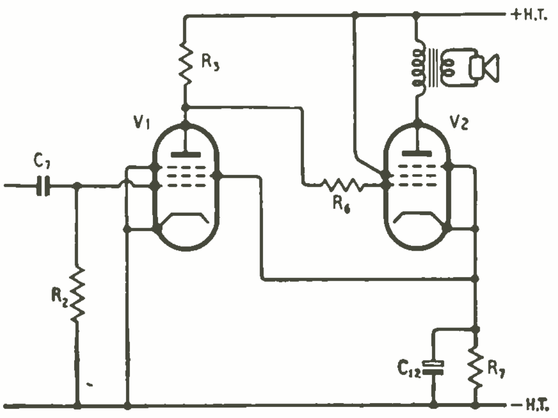


Fig. 1. Basic circuit of the receiver illustrating the starvation technique adopted.

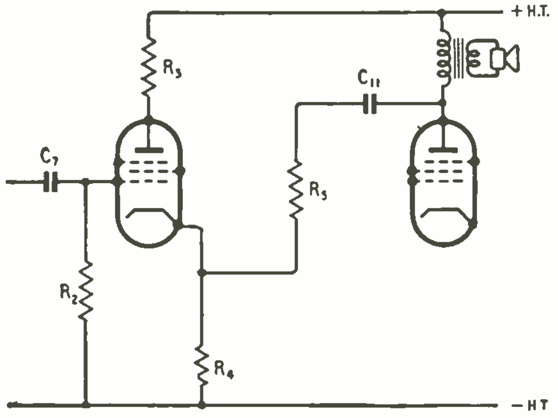


Fig. 2. Method of applying negative feedback without risk of instability.

Receiver

By S. W. AMOS,* B.Sc. (Hons.) A.M.I.E.E.

circuit. Practically the only form of gain control which can be used is one employing variable negative feedback. Such a method of gain control has the advantage that any gain in excess of that required at any moment is not " thrown away " in an attenuator, but is usefully employed in reducing distortion and improving loudspeaker damping.

Feedback gain controls usually have a number of disadvantages; for example their range is often inadequate, because the maximum degree of feedback is limited to a value which does not cause instability, and output volume cannot be reduced to zero. The circuit used in this receiver is free from these limitations; it was developed from the circuit shown in Fig. 2 in which a feedback potentiometer R_4R_5 is connected between the anode of V2 and the cathode of V1. This arrangement permits a very large degree of negative feedback without instability; in fact R_5 can be reduced to zero without provoking oscillation.

To give control of gain either R_4 or R_5 can be made variable. If R_4 is variable, it must have an inverse logarithmic law to give smooth control of volume; on the other hand if R_5 is variable this must have a logarithmic law to give smooth gain control. Accordingly R_5 is made variable and the circuit takes the form shown in Fig. 3. A further advantage of making R_5 variable is that, at the position of minimum gain, it effectively short-circuits the primary winding of the output transformer to give zero output from the receiver. However, the shunting effect of R_5 on the primary winding is undesirable at settings of R_5 other than near the minimum. This can be minimized by choosing the values of R_4 and R_5 in the following way. For reception of a reasonably strong signal the a.f. gain of the receiver averages approximately 1000. For such values of gain, the gain is determined by the constants of the feedback loop and is given approximately by R_5 R_4. To keep the a.f. loss in R_5 reasonably

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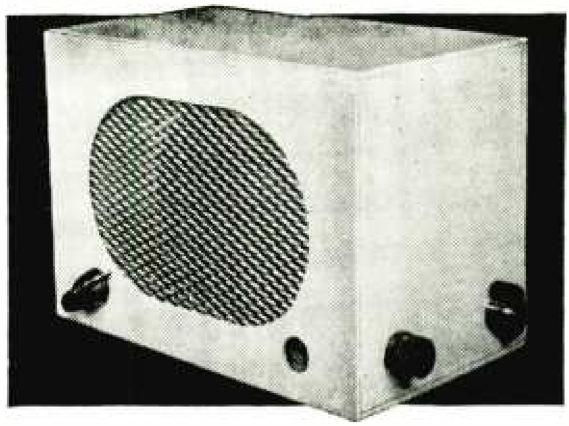


Translucent Perspex sheet was used for making the cabinet.

low, R_5 should preferably not be less than 100 kΩ (5 times the effective loudspeaker impedance at the primary winding). This gives the minimum value of R_4 as 100 Ω and a value of 140 Ω is used.

Maximum Gain

There now arises another difficulty. To obtain maximum gain from the receiver there should be no feedback when R_5 is set to its maximum value. This requires that the ratio of R_5 to R_4 should be large compared with the internal gain of the circuit (i.e., the gain in the absence of feedback). The internal gain is approximately 60,000 (150 from V2 and 400 from V1) and thus R_5/R_4 should preferably not be less than say, 300,000. Since R_4 is 140 Ω, R_5 must be 42 MΩ! There is, however, no need for such a large value if the " free " end of R_5 is returned to h.t. negative as shown in Fig. 4. When R_5 is advanced to its maximum setting it now short-circuits R_4 thus removing feedback and giving maximum gain. R_5 can be a standard logarithmic volume control of 1 MΩ.

Finally the value of C_{11} must be determined. This must be fairly large because R_5 may be 10 kΩ or less at low volume settings and, if the reactance of C_{11} is comparable with this at low audio frequencies, an

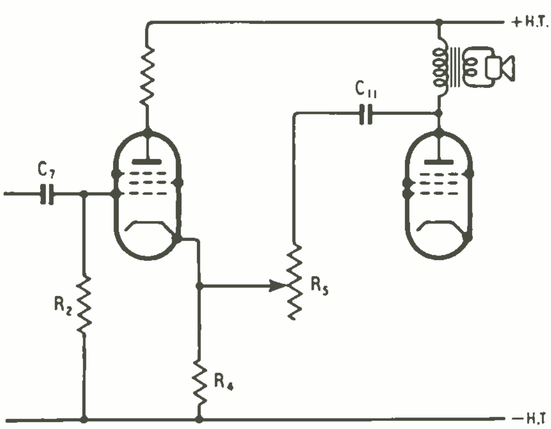


Fig. 3. First stage in the development of the feedback gain control.

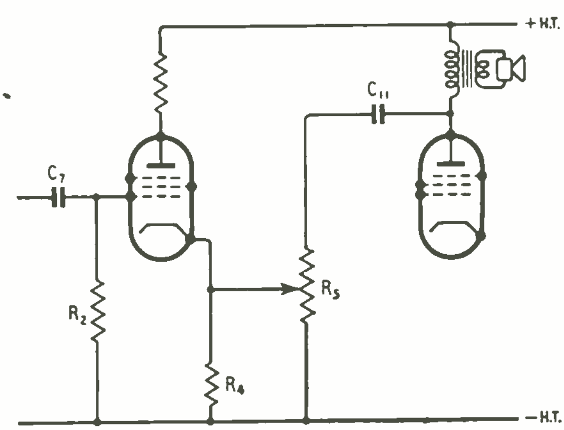


Fig. 4. Final circuit diagram of the feedback gain control giving maximum volume range.

accentuation of the lower audio frequencies results. To restrict any bass lift to less than 1 db at 50 c/s when R_5 is 10 kΩ, C_{11} must be greater than 0.6 μF and a miniature 4-μF electrolytic capacitor is used.

To keep the receiver simple it was decided to use only a single LC circuit for tuning. Such a simple tuning arrangement is, of course, incapable of giving good reception of weak signals when there are strong ones on the same waveband but, with the aid of reaction, it has proved capable of separating the two medium-wave signals in the London area without pressing reaction to the point of oscillation. For ease of operation it was decided to employ pre-set tuning, the Home or Light programmes being selected by a 2-way switch. This decision simplified the problem of coupling the aerial to the tuning circuit, for it is possible to connect the aerial to the " hot " end of the LC circuit *via* a fixed capacitor as shown in Fig. 5. Such simple coupling can be very effective in a circuit operating at fixed frequency but is unsatisfactory in receivers with variable tuning because gain and selectivity are greatly dependent on frequency and vary considerably over the band.

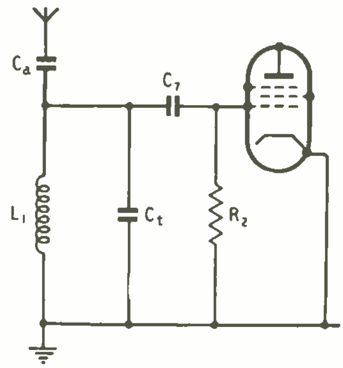


Fig. 5. Method of aerial coupling used in the receiver.

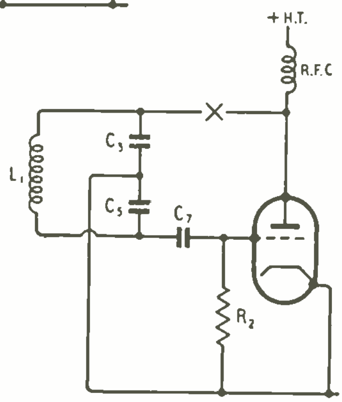


Fig. 6. Colpitts oscillator circuit from which the reaction circuit was derived.

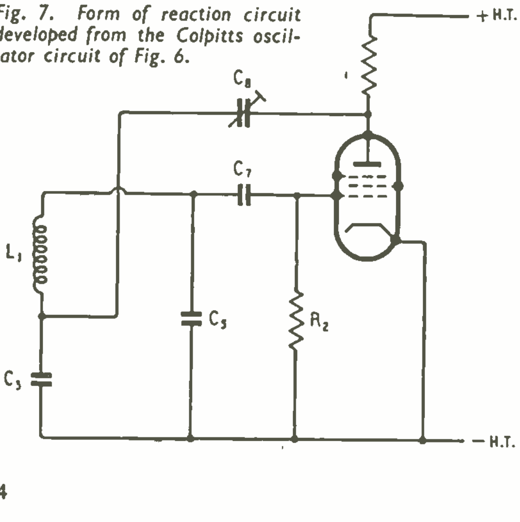


Fig. 7. Form of reaction circuit developed from the Colpitts oscillator circuit of Fig. 6.

As shown in the appendix, the gain of an aerial coupling circuit of this type is given approximately by C_aQ/C_t where C_a is the coupling capacitance, C_t is the tuning capacitance and Q is the reactance/resistance ratio of the inductor L_1. Thus if C_a is an appreciable fraction of C_t, the gain is an appreciable fraction of Q. For example if C_a = 50 pF and C_t = 200 pF a gain equal to Q/4 is available. This is hardly a practical condition, however, for if C_a is 50 pF, the effective tuning capacitance is greatly affected by variations in aerial capacitance. Thus the calibration of the receiver tuning control is dependent upon the aerial constants and varies from aerial to aerial. It is particularly desirable that the tuning of the receiver should be substantially unaffected by changes in aerial constants, because adjustment of tuning is not so convenient as in a receiver without pre-set tuning. This condition can be achieved by making C_a small compared with the capacitance of the aerial itself; a value such as 10 pF is suitable. To achieve high gain with C_a = 10 pF, C_t must also be small, say 30 or 40 pF. To tune the Home Service (1088 kc/s in the Midland area, for which this receiver was destined) with such a small capacitance necessitates an inductor of approximately 700 μH. Such a value was accordingly used. It enables the whole of the medium waveband to be covered with a capacitor of 120 pF maximum capacitance, and also has the advantage that the long-wave Light Programme (200 kc/s) can be tuned with a capacitor of 1000 pF.

Thus the medium-wave Home Service and the long-wave Light Programme are both tuned using the same inductor by simple selection of capacitors. The coil used (Osmor QIF1) is a standard component used in a 465 kc/s i.f. wavetrap and gives an inductance range of 500 to 800 μH by adjustment of the iron core. This inductance adjustment is used for tuning long waves (the 1000-pF capacitor being fixed) and the Home Service is tuned with an adjustable trimmer of 70 pF maximum capacitance. Since C_t is 1000 pF on long waves C_a should be approximately 200 pF to maintain the same gain as on medium waves. This is of the order of the capacitance of an aerial and thus no physical capacitance is necessary for long-wave coupling, the aerial being connected directly to the top of the coil. Variations in aerial capacitance are unlikely to affect the long-wave calibration because they will in general be small compared with the 1000-pF tuning capacitance.

Reaction Circuit

Reaction is usually obtained by use of an additional inductor closely coupled to the tuning inductor, but such a circuit is not suitable for use in this receiver because it requires a coil assembly which would need to be specially wound. It was decided therefore to use a reaction circuit which can be applied to a single untapped inductor such as that chosen for tuning purposes. The circuit adopted is derived from that of the Colpitts oscillator shown in its usual form in Fig. 6. A significant feature of this circuit is that oscillation is most vigorous when C_3 is equal to C_5, the effective tuning capacitance being then $C_3/2$. The amplitude of oscillation can be controlled by a variable capacitor C_8 introduced at the point X and if this capacitance is reduced below a certain value, oscillation ceases,

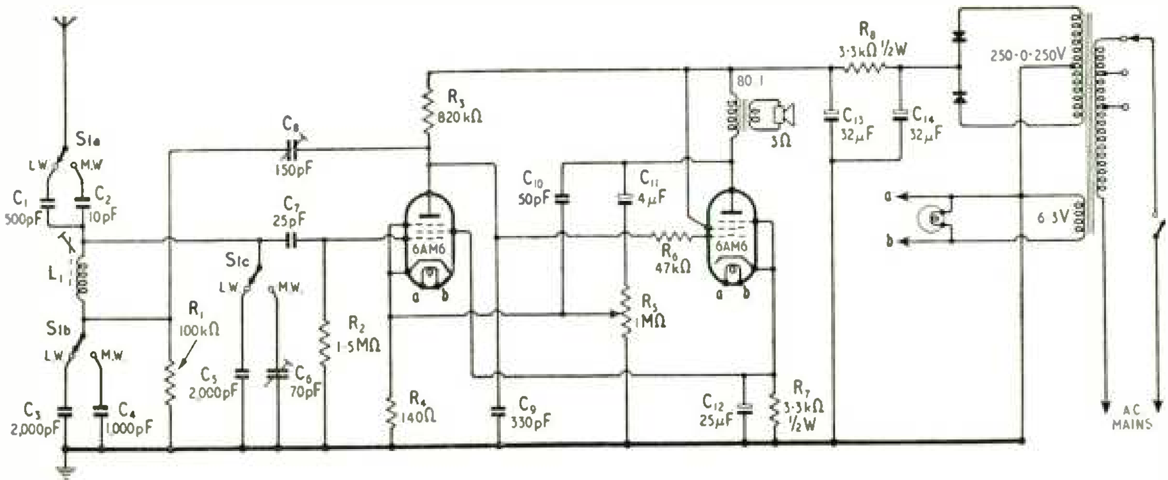


Fig. 8. Complete circuit diagram of the receiver for reception of one medium-wave and one long-wave signal. All resistors can be ¼ W unless otherwise specified.

the circuit then resembling that of a detector in which C_8 acts as a reaction control. This is precisely the circuit used for long-wave reception although in the complete circuit diagram it is drawn in the form shown in Fig. 7. C_3 and C_5 are both 2000 pF, giving the required effective tuning capacitance of 1000 pF. A maximum value of 150 pF is adequate for C_8 and adjustment of its value has no significant effect on tuning. The fact that C_3 and C_5 are both equal implies that only one half the signal developed across L_1 is applied to the detector.

As shown in the complete circuit diagram (Fig. 8), for medium-wave reception C_5 (2000 pF) is replaced by C_6 (70 pF maximum) and C_3 (2000 pF) by C_4 (1000 pF). If C_4 and C_6 were made equal C_8 would need to be very much smaller to control reaction on medium waves than on long waves. In a pre-tuned receiver it is desirable that the reaction control should not require readjustment after each operation of the station-selector switch. The value of C_4 is chosen to satisfy this condition as far as possible but the capacitance of C_8 required to give oscillation on medium waves increases as C_6 is decreased; this differs from the behaviour of the more usual reaction circuits. The large ratio of C_4 to C_6 ensures that nearly all the signal developed across L_1 is applied to the detector and it also enables the whole of the medium waveband to be covered by variation of C_6 alone. To cover the whole of the band C_6 should be 120 pF maximum; a value of 70 pF was used by the author to tune in the Midland Home Service. Medium-wave

TABLE

Maximum wavelength in metres	Minimum frequency in kc/s	Maximum capacitance of trimmer required in pF
250	1200	25
300	1000	35
350	857	50
400	750	65
450	667	80
500	600	100
550	545	120

These values are based on an inductance of 700 μH

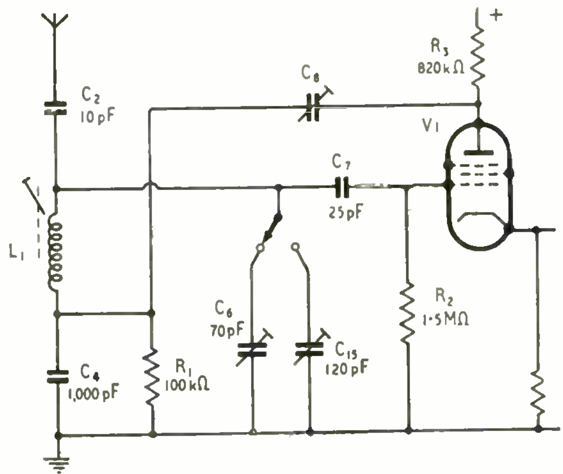


Fig. 9. Input circuit suitable for reception of two medium-wave stations.

tuning is substantially unaffected by operation of the reaction control.

A third section of the station-selector switch connects a 10-pF capacitor (C_2) in the aerial lead for medium-wave reception. If long-wave reception is not required there is no need for the 2000-pF capacitors or to change the series aerial capacitance and C_4 can be retained for both stations as shown in Fig. 9. As suggested in this circuit C_6 could be, say, 70 pF for a station near the high-frequency end of the band and C_{15} could be 120 pF maximum for a station at the other end. If the receiver is pre-tuned to two medium-wave signals of widely different frequencies, say one near 600 kc/s and the other near 1.5 Mc/s it may be desirable to have different values of C_4 for the two signals to give approximately the same degree of regeneration on both. A value of C_4 of 2000 pF is suitable for 600 kc/s and 500 pF for 1.5 Mc/s, a value of 1000 pF being suitable for 1 Mc/s as shown in Fig. 8. Approximate values of C_6 necessary for various frequencies are given in the table.

R_1 and C_1 are included in the circuit to provide some attenuation to 50-c/s signals from the aerial.

Without R_1 the tuning inductor has a high impedance to earth at 50 c/s and on long waves, when the aerial is in direct contact with L_1, 50 c/s signals can be of sufficient amplitude to modulate a received signal. This trouble is not present to any extent on medium waves because of the very high reactance at 50 c/s of the series capacitor C_2.

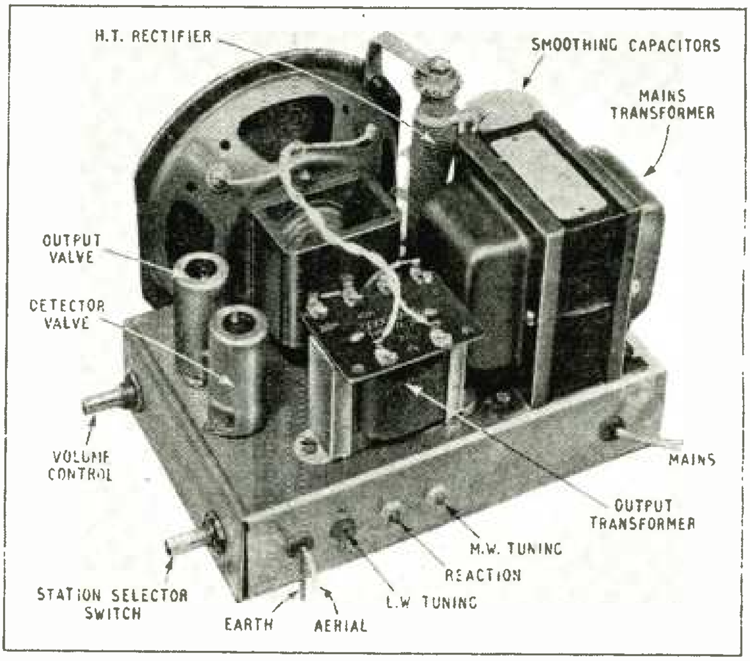
As mentioned earlier, in certain circumstances, results from a particular transmission may be unsatisfactory due to overloading of the detector. If this occurs the input can be reduced by decreasing the appropriate series input capacitor (C_1 or C_2).

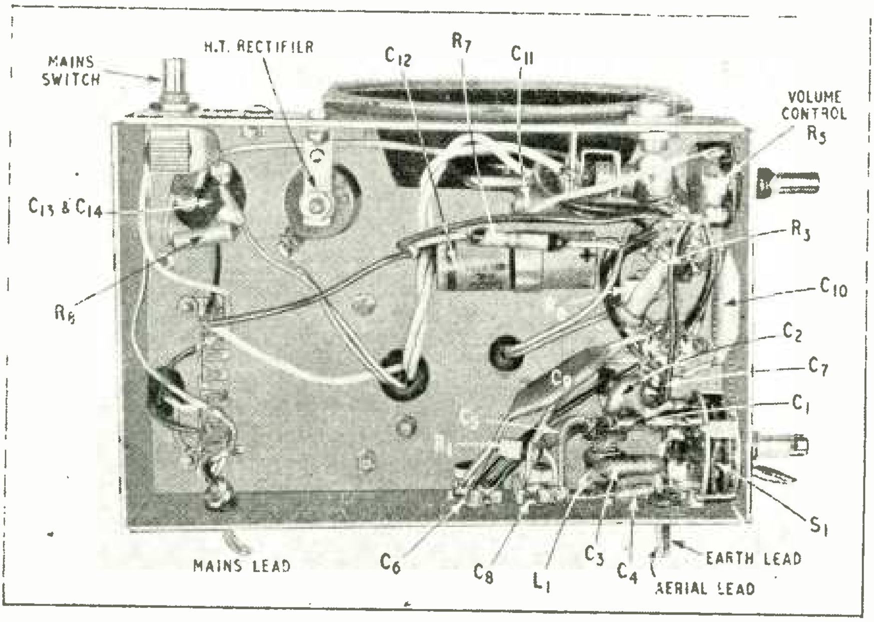
In receivers with pre-set tuning it is advantageous to have equal volume from all stations; the station-selector switch can then be operated without necessity for subsequent volume readjustment. This can be achieved in this circuit by correct choice of values for C_1 and C_2. If both transmissions give good signals, one being stronger than the other, the capacitor corresponding to the stronger of the two signals can be decreased until the volume obtained is equal to that from the other signal.

The only other points of note in the circuit are the capacitors C_9 and C_{10} which are for r.f. decoupling and the resistor R_6 which also provides some r.f. attenuation with the input capacitance of V2 but which is primarily intended as a grid stopper.

A full-wave selenium rectifier is used to supply h.t. and the smoothing circuit $R_7C_{13}C_{11}$ supplies 280 volts at 12 mA. Approximately 40 volts are lost across R_7 and the effective h.t. supply for V2 is thus 240 volts.

The construction of the receiver is illustrated in the accompanying photographs. The four-sided chassis measures 7 inches by 4½ inches by 1½ inches and much of it is occupied by the mains transformer which is rated at 250-0-250 volts 60 mA and has a single 6.3-volt l.t. winding. This is an unnecessarily generous rating for such a small receiver but this type of transformer (Electro-Voice Type 104E) is used because it is fairly small and readily obtainable. The selenium rectifier is a 250-volt 60-mA bridge type, a government surplus component type 280 LU997AW used here as a push-pull rectifier after removing the link joining the outermost tags. The current rating is again unnecessarily high and any push-pull rectifier rated for 250 volts and capable of supplying 12 mA will be satisfactory. The output transformer is a Goodmans Type 74/243. The station-selector switch and volume control are mounted on one end wall of the chassis, the on-off switch and





Plan view of chassis is shown above. The mains transformer is somewhat larger than it need be. On the left is the underside of chassis showing positions of most resistors and capacitors.

indicator lamp on the front, tuning and reaction controls being on the rear flange.

In setting up the receiver it is necessary to tune in the long-wave programme first, by adjustment of L_1, and the medium-wave programme afterwards by adjustment of C_6. If possible the 6AM6 used for detection should be specially selected because some valves of this type tend to be microphonic and can set up continuous oscillation by acoustic feedback from the loudspeaker.

The cabinet illustrated was home-made of Perspex, and the internal dimensions are 8½in × 6in × 5in. The length of the chassis is thus 1½ inches less than the corresponding dimension of the cabinet, this margin being necessary when the chassis is inserted in the cabinet to enable the controls on the end wall of the chassis to be fitted into the corresponding holes in the cabinet.

APPENDIX

The essential features of the aerial-coupling circuit are shown in Fig. 10 in which r and c represent the resistance and capacitance respectively of the aerial-earth system. Values of r and c commonly used in medium-wave dummy aerials are 40 ohms and 200 pF. C_a is the coupling capacitor and r_L is the r.f. resistance of the inductor L_1.

Maximum voltage is developed across C_t at the frequency at which the net inductance of L_1, C_t and r_L resonates with the net capacitance C of c and C_a in series.

The impedance Z of the network L_1, C_t and r_L is given by

$$Z = \frac{(j\omega L_1 + r_L)/j\omega C_t}{j\omega L_1 + r_L + 1/j\omega C_t}$$

$$= \frac{j\omega L_1 + r_L}{1 + j\omega C_t r_L - \omega^2 L_1 C_t}$$

By rationalizing this expression we can show that the network is equivalent to a series circuit of inductance given by

$$L = \frac{L_1}{1 - \omega^2 L_1 C_t} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

and resistance given by

$$R = \frac{r_L}{(1 - \omega^2 L_1 C_t)^2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

Since the inductance (1) resonates with the capacitance C we have

$$\frac{1}{\omega C} = \frac{\omega L_1}{1 - \omega^2 L_1 C_t}$$

from which

$$1 - \omega^2 L_1 C_t = \omega^2 C L_1$$

Substituting for $(1 - \omega^2 L_1 C_t)$ in (2)

$$R = \frac{r_L}{\omega^4 C^2 L_1^2}$$

Thus the circuit is equivalent to the simple series network shown in Fig. 11. The voltage gain of this circuit is equal to the quotient of the reactance (of the inductance or the capacitance) and the resistance, thus

$$\text{Gain} = \frac{1/\omega C}{r + r_L/\omega^4 C^2 L_1^2}$$

If practical values are substituted for the symbols in this expression it is found that r is normally small

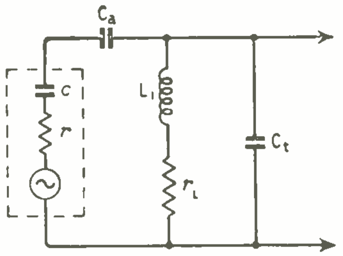


Fig. 10. Aerial-coupling circuit with aerial replaced by equivalent generator.

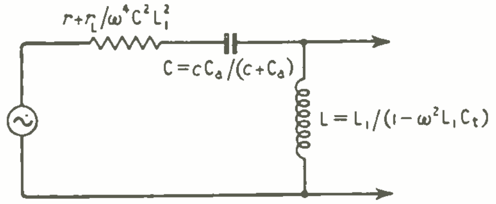


Fig. 11. Simple series circuit equivalent to the network of the previous figure.

in comparison with the other term in the denominator. If r is neglected the gain is given approximately by

$$\text{Gain} = \frac{\omega^3 C L_1^2}{r_L}$$

Now r_L is given by $\omega L_1/Q$ and substituting for r_L we have

$$\text{Gain} = \omega^2 C L_1 Q$$

which shows that the gain is proportional to the square of the frequency. It thus varies in the ratio 9 : 1 over the medium waveband.

C_a is made small compared with c in order to make the calibration of the receiver substantially independent of variations in c. For such values of C_a the frequency of maximum gain is approximately the resonance frequency of L_1 and C_t. Thus $\omega^2 L_1$ may be replaced by $1/C_t$ and we have

$$\text{Gain} = \frac{CQ}{C_t}$$

If C_a is small C is approximately equal to C_a. Hence

$$\text{Gain} = \frac{C_a Q}{C_t}$$

which is the result used in the text.

"DECADE COUNTER"

Correction

An error unfortunately appeared in this article in the May issue. In Fig. 2 the second feedback path should go from the anode of V2 in the fourth stage to the grid of V1 in the third stage. Then, in the right-hand column of page 235 the section beginning 20 lines from the top, "This time, a negative pulse . . ." should be deleted to the end of the paragraph and be replaced by the following: "The third stage remains in this condition (the original state) for only a fraction of a microsecond because this transition causes a reversal of itself. The negative transient produced when the third stage is triggered by the second stage switches stage four through a half-cycle of its operational cycle. A positive pulse is thereby returned to point "C" of the third stage, re-triggering it extremely quickly. So brief was the excursion of the third stage to its original state that insufficient pulse energy is delivered to the second stage via the feedback path to cause any disturbance."

Recording Low Frequencies on Magnetic Tape

By D. W. THOMASSON, A.M.Brit., I.R.E.

The Application of Pulse Code Techniques

NOT many years ago it used to be said that the upper frequency limit of a tape recording system was approximately one kilocycle per inch of tape speed. Today, with responses sensibly flat to 15 kc/s at a speed of 7½ inches per second, this is no longer true. There has been no comparable development at the other end of the frequency range, however, the "hum barrier" at 50 c/s still setting a limit to low frequency response in most cases.

The difficulty arises from the small output given by the playback head at low frequencies, typical figures being 200μV at 50 c/s and 100μV at 25 c/s. Since 1.5μV effective input hum represents good performance in an amplifier with an a.c. heater supply, the signal/hum ratio cannot exceed 40 db, and even when a d.c. heater supply is used hum pick-up imposes a serious limitation. At lower frequencies the problem of obtaining high gain without instability sets the ultimate limit.

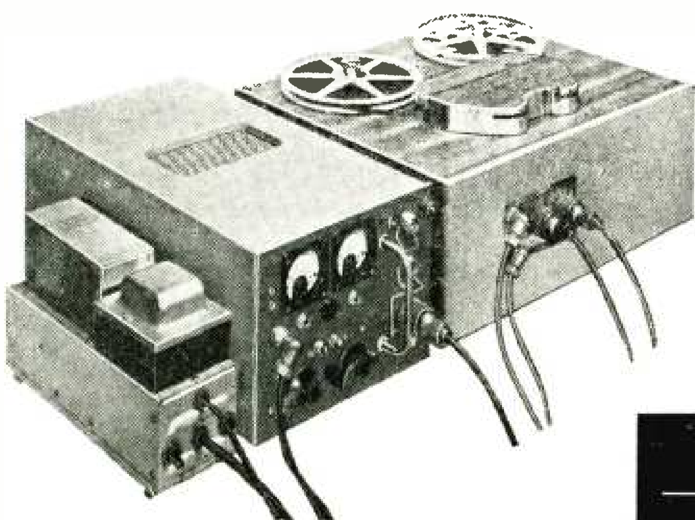
For recording the very low frequencies involved in some types of scientific and industrial measurement, a pulse code method can be used. Two systems are employed in apparatus introduced by Messrs. Rudman, Darlington (Electronics), Ltd., Clyde Works, Lichfield Road, Wednesfield, Wolverhampton, Staffs., one covering the 0-150 c/s range with manual setting of the d.c. level (tape speed 3¾ inches per second) and the other covering the 0-250 c/s range with fully automatic d.c. maintenance (tape speed 7½ inches per second).

The first method uses pulse interval modulation, Fig. 1 (a). The pulses are of uniform length and amplitude, varying only in their separation, and they are converted by a charging circuit into triangular pulses having a mean level that follows the modulation. The coding frequency component is filtered out, leaving the original input signal.

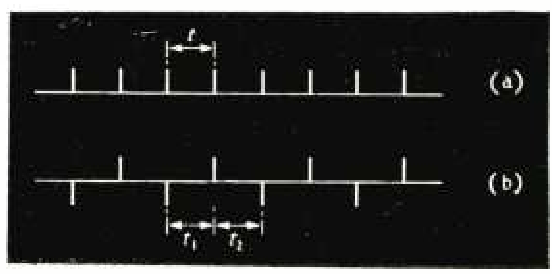
It will be appreciated that, while the system is unaffected by a reasonable degree of hum and noise superimposed on the code signal, any variation in tape speed alters the d.c. level at the output, and provision is therefore made to balance out any spurious d.c. component by manual adjustment. A very good performance in respect of wow and flutter is essential and the tape drive used has a maximum variation of only 0.2 per cent at 3¾ inches per second, giving a negligible disturbance of the output signal. A second recording channel on the same tape deals with higher frequencies up to 6 kc/s using orthodox techniques.

While this method of coding allows maximum tape economy, it was recognized that there is a need for a system in which the d.c. component is maintained without manual adjustment, and the second method meets this need. In this case all low-frequency stages are of balanced d.c.-coupled form, allowing a higher input sensitivity, and the coded signal conveys two information channels, each represented by a set of time intervals, Fig. 1 (b). With zero input voltage all the time intervals are equal and the twin decoders pass equal voltages to the balanced output. With any other input voltage level one interval is increased, the next decreased, and so on, the decoder signals becoming unequal by a corresponding amount. The output is proportional to the difference in the two time intervals divided by their sum, and since all the intervals change in

(*Continued on page* 549)



Twin-channel pulse coded strain gauge magnetic recorder, for use in aircraft.



Right: Fig. 1. (a) Simple pulse time system (b) push-pull system giving modulation amplitude independent of tape speed.

the same proportion when the tape speed is altered the recording can be played back at any speed without altering the output voltage levels.

The more complex waveform of the coded signal makes heavier demands on the actual recording process and it is interesting to note that the success of the system largely depends on the improvement in high frequency response which has already been mentioned.

Both methods are of especial interest in strain gauge and vibration testing, but can be applied in many fields where pen recorders are usually employed. Recordings can now be inspected and selected before transfer to paper and the time scale can be changed by alteration of the tape speed if it is necessary to ease the demands made on pen recorder performance.

Another interesting application arises in connection with the flight testing of aircraft. The equipment illustrated is used for flight testing jet engines, and is designed to minimize the effect of the aircraft's movements. If the recorder cannot be mounted in the aircraft the second coding system can be used to provide a telemetering system that is unaffected by the signal fading which is inevitable over air-to-ground paths. The technique is likely to be of even greater interest soon, as current development work is aimed at recorders giving four or eight independent channels on a single tape.

LEIPZIG FAIR

Impressions of Eastern European Radio Products

By V. A. SHERIDAN, A.M.I.E.E.*

COMMENCING with the ordinary domestic radio receivers, these were being exhibited in profusion by most East European countries. In appearance and performance they are not unlike ours. Radio-gramophones, however, were few and far between and certainly had no automatic record changers. Generally, all receivers are of the table type, housed always in wooden cabinets.

Of television receivers, I only saw the East German product being demonstrated. The sets are of the table type and are fitted with an approximately 8-inch tube The brightness is very poor and the picture can only be viewed in complete darkness. Also there is a distinct flicker present obviously due to insufficient persistence of the screen material. The definition is good, as would be expected from the 625-line system. The only other exhibitors of television receivers were the U.S.S.R. who are showing 9-inch table models. These sets were not being demonstrated.

The German radio receivers are priced between 300 and 400 marks. It is difficult to translate this into our currency as the official rate of exchange is 6.20 marks to the pound. I think the West German rate of 12 marks to the pound does give a fair comparison in relation to the people's income. Television sets are priced at 1,300 marks. Needless to say, the popular demand for these sets is negligible.

The valves are much of the same design as ours; that is the all-glass construction with bases similar to our B7G and B9A. The main difference is that the contact wires are fitted with shaped sleeves soldered on to the wires. This results in a rather more positive grip in the valve holder itself and dispenses with valve retainers. Specialized valves such as magnetrons and klystrons, etc., were also exhibited. Crystal diodes and transistors were conspicuous by their absence.

Cathode-ray tubes displayed were of the all-glass design as well as the metal-glass construction. A wide range of oscilloscope tubes of the single and double-beam variety were exhibited.

In the component field a very wide range was shown. The makes known before the war are being produced under new names, as practically all firms have been nationalized, and are called People's Owned Works.

The general range and design is much the same as ours. However, nearly all paper capacitors employ the metallized paper process, resulting in very neat and small units. Carbon resistors are all of Grade I type and are made from 1/20th watt to 2 watts. I have brought back some samples and find they are equal to the best British ones whilst their price is about 1/5th of what we pay here. Furthermore, they supply a precision-type high stability resistor with 0.5 per cent limits. For high resistance values they have a so-called "colloidal" type. I was assured by the chief research engineer of one of the firms concerned that they do not employ a colloidal graphite coating. The standard values are up to 10^{10} ohms, whilst small quantities can be supplied up to 10^{12} ohms. They claim that a maximum voltage of 1,000 volts d.c. can be applied. The resistors are contained in evacuated glass envelopes.

A wide range of rotary switches are available. The makers of one type claim a maximum contact resistance of 20 milliohms. I confirmed this value on a sample. However, by applying a contact oil the value dropped to 5 milliohms. This value was maintained after 3,000 operations.

Components which employ much metal are not of the standard which we are accustomed to. This is due to the acute shortage of all raw materials in Eastern Germany. For instance, brass will be used instead of copper wherever possible, in order to save the copper. It must be borne in mind that the whole economy is working under typical wartime conditions throughout.

Regarding instrumentation, not only did the East Germans show a wide range, but also the Russians, the Czechs and the Hungarians. The equipments cover the entire range from indicating instruments up to equipment for the measurement of centimetre waves.

The general appearance and finish of the East German instruments is very good. In particular they have developed excellent designs of dials, resulting in a very clear indication. The dial rotates according to the multiplier setting, permitting up to eight ranges, thus completely eliminating reading errors.

* British Physical Laboratories.

Some Electrical Theorems

Their Practical Utility

By W. TUSTING

TEXTBOOKS of the more mathematical kind abound in theorems having more or less high-sounding names. The ordinary man is apt to pass over such matters as being difficult and of little practical value. In this, however, he is quite wrong, for some of them are not at all hard to understand or to remember and they are of considerable utility. Apart altogether from their mathematical applications they sometimes help considerably to an understanding of circuits.

Thévenin's Theorem

One of the best-known of such theorems is the one commonly known as Thévenin's theorem. This states that any linear network, no matter how complex, containing any number of sources of e.m.f. is, when regarded from any pair of terminals, equivalent to an impedance in series with an e.m.f.; the impedance is that measured between those terminals with all internal sources of e.m.f. short-circuited and the e.m.f. is the open-circuit voltage at those terminals.

This sounds very difficult, but a few simple examples will make it clear. Suppose the network comprises a potential divider R_1 and R_2 connected to a battery E as shown in Fig. 1. The impedance measured between the terminals with E short-circuited (if one were making a real measurement instead of an imaginary one, one would naturally remove the battery and short-circuit the terminals to which it had been connected, just to avoid destroying the battery!) is clearly R_1 and R_2 in parallel. Call it

$$R = \frac{R_1 R_2}{R_1 + R_2}$$

The voltage measured across the terminals on open-circuit (that is, with no current-consuming device connected to them) is plainly

$$V = E\frac{R_2}{R_1 + R_2}$$

The equivalent circuit is thus Fig. 2(b) and it is in every respect identical in performance with the more complex original (a).

A numerical example may help here. Suppose E is 250 V, the h.t. line of a receiver, and R_1 and R_2 form a potential divider to provide a lower voltage supply; suppose R_1 is 100 kΩ and R_2 is 25 kΩ. Then $V = 250 \times 25/125 = 50$ V, and $R = 100 \times 25/125 = 20$ kΩ. The supply obtained in this way is exactly the same as one obtained from a 50-V source through a 20-kΩ resistor.

The theorem holds for a.c. as well as d.c., but the voltage may then become frequency dependent. Consider Fig. 2(a) in which a resistor R and a capacitor C are connected to an a.c. generator *e*. Applying Thévenin's theorem gives (b) the internal impedance being R and C in parallel. The generator voltage is

$$v = e\frac{1/j\omega C}{R + 1/j\omega C}$$

and this varies in magnitude and phase with the frequency.

The theorem is quite valid under this condition but is less useful. In this particular example there is really no point in using the theorem at all, for it tends to complicate matters rather than to simplify them. However, with a circuit like Fig. 3(a) its use is very helpful if it is applied discriminatingly.

The thing to do here is to disregard C for the time being. Then apply the theorem to *e* and the resistors only. This bit of the circuit is the same as Fig. 1(a) and has the equivalent of Fig. 1(b). We now put back

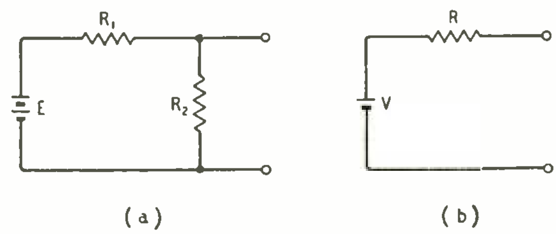


Fig. 1. Simple potential divider and battery (a) and its Thévenin equivalent (b).

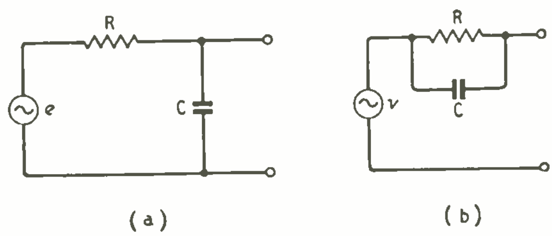


Fig. 2. An a.c. generator with an RC circuit (a) can be transformed to (b) but in this case there is rarely much advantage in doing so.

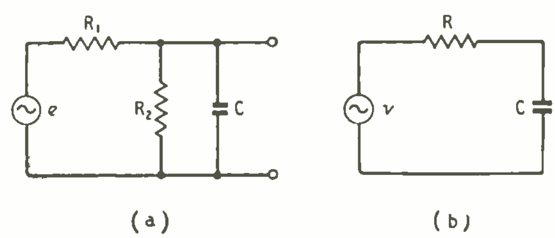


Fig. 3. An RC circuit of form (a) can be advantageously transformed to (b).

the capacitor and get Fig. 2(a) and draw it again as Fig. 3(b) where

$$v = e\,\frac{R_2}{R_1 + R_2} \quad \text{and} \quad R = \frac{R_1R_2}{R_1 + R_2}$$

The theorem can be applied in reverse. In Fig. 4(a) is shown a resistance-coupled stage, which might be the sync separator of a television receiver. The required value of R_a might be 10 kΩ and we might wish to operate the stage from a 20-V supply whereas the h.t. line might be 200 V. The natural thing to do is to make R_a = 10 kΩ and to obtain the 20-V supply by a potential divider R_1R_2 as in Fig. 4(b) and this is necessary if decoupling as provided by C is needed. If it is not, we can leave off C. We can then apply Thévenin's theorem to R_1R_2 and get Fig. 4(c) and we can see that to keep the load on the valve at 10 kΩ we shall have to reduce R_a by the value of R.

It now becomes obvious that there is a redundant resistor, for if R is made 10 kΩ and V is made 20 V, R_a can be dispensed with and the potential divider itself becomes the load. This is shown at (d). For our figures we have

$$20 = 200\,\frac{R_2}{R_1 + R_2} \quad \text{and} \quad 10 = \frac{R_1R_2}{R_1 + R_2}$$

so

$$\frac{R_2}{R_1 + R_2} = \frac{1}{1 + R_1/R_2} = \frac{1}{10} = \frac{10}{R_1}$$

and

$$R_1 = 100\ \text{k}\Omega,$$

$$R_1/R_2 = 9, \quad R_2 = 11.1\ \text{k}\Omega$$

This form of circuit, in which the load and voltage-droppers are combined in potential-divider form is sometimes used in television receivers. It is a bit puzzling when first met but is easily unravelled with the aid of Thévenin's theorem.

Norton's Theorem

One could go on quoting examples of the application of Thévenin's theorem indefinitely but enough has been said to show its utility and the time has come to turn to another—Norton's theorem. This is a very similar one and states that any network, containing any number of sources of e.m.f. is, when regarded from any pair of terminals, equivalent to an impedance in shunt with a current generator of infinite internal impedance; the impedance is that measured between those terminals with all internal sources of e.m.f. short-circuited and the current is the current which will flow between the terminals when they are short-circuited.

Let us apply this to the circuit of Fig. 5(a), which is the same as Fig. 1(a). The impedance measured at the terminals is

$$R = \frac{R_1R_2}{R_1 + R_2}$$

as before, and the equivalent circuit is Fig. 5(b) where I is the current generator. The short-circuit current in (a) is E/R_1 and this is the value of I in (b).

Norton's theorem is much less used than Thévenin's in this general sense, but it is very widely used in connection with pentode valves. The ordinary equivalent circuit of a valve is of the form of Figs. 1(b) or 3(b) and is expressed like Fig. 6(a). By the use of Norton's theorem it can be put in the form of Fig. 6(b) which is equally known. The resistance is the same in both. In (a) the short-circuit current is $\mu v_g/r_a = g_m r_a$, which is the current generator of (b). It is only because we commonly write μ/r_a as g_m, the mutual conductance, that we do not always recognize (b) as a transformation by Norton's theorem of (a).

Star-Delta Theorem

Another very useful theorem is that commonly known as the star-delta theorem, but also called the T-delta, T-Δ, or T-π theorem. Any three impedances in the

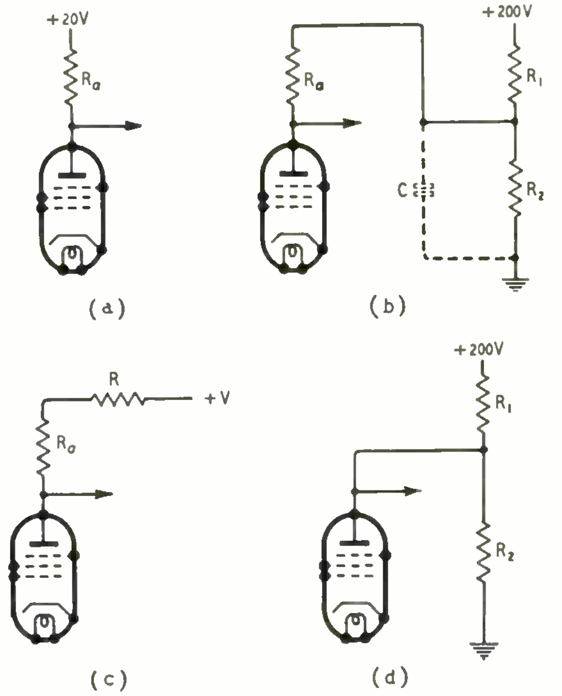


Fig. 4. A valve with a load R_a which requires a low h.t. voltage as in (a) might be used with a potential divider R_1R_2 when the supply is of high voltage. The Thévenin transformation is (c) and shows there is an unnecessary resistance and so the final circuit can be reduced to (d).

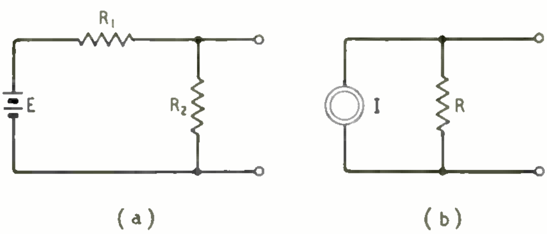


Fig. 5. By Norton's theorem these two circuits are equivalent.

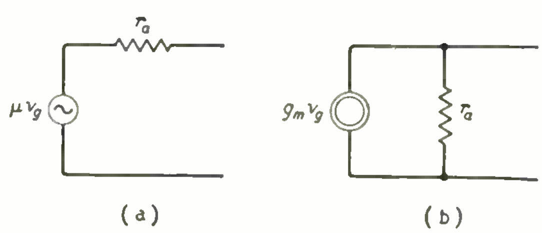
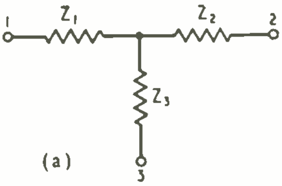
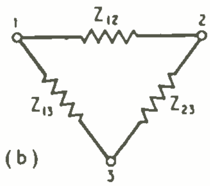


Fig. 6. Norton's theorem in the case of a valve: (a) is the circuit commonly used to represent a triode and (b) that used for a pentode.



$$Z_1 = Z_{12}Z_{13}/(Z_{12}+Z_{13}+Z_{23})$$
$$Z_2 = Z_{12}Z_{23}/(Z_{12}+Z_{13}+Z_{23})$$
$$Z_3 = Z_{13}Z_{23}/(Z_{12}+Z_{13}+Z_{23})$$



$$Z_{12} = Z_1 + Z_2 + Z_1Z_2/Z_3$$
$$Z_{13} = Z_1 + Z_3 + Z_1Z_3/Z_2$$
$$Z_{23} = Z_2 + Z_3 + Z_2Z_3/Z_1$$

Fig. 7. The star and delta networks shown here are identical if the impedances have the relations shown.

star or T form of Fig. 7 (a) can be transformed into three different impedances in delta or π form as shown at (b) or, of course, vice versa. The relations between the impedances are given in the figure. If the network is symmetrical (i.e., $Z_1 = Z_2$ or $Z_{13} = Z_{23}$) the relations simplify considerably.

There are many uses for this equivalence. One simple one is in attenuators, perhaps for a television aerial feeder. In such a case, the values for a star might be $Z_1 = Z_2 = 59\ \Omega$ and $Z_3 = 14.6\ \Omega$, values which would give 20-db attenuation for a feeder impedance of 72 Ω. It might well happen that one had no suitable resistors available and the delta equivalent might be more convenient. From Fig. 7,

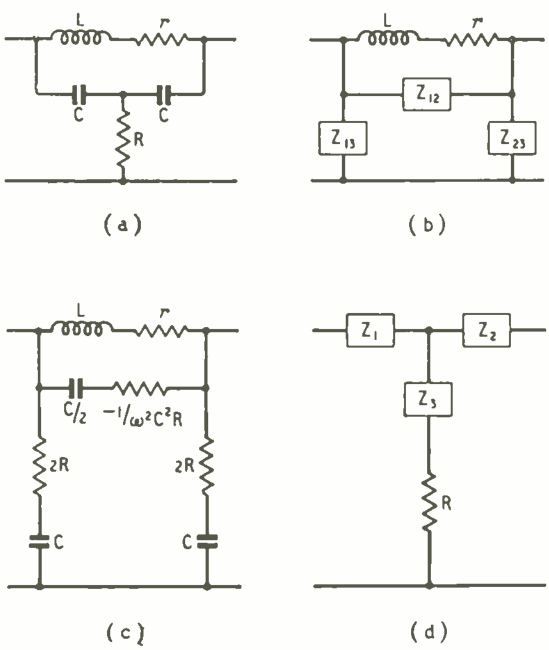


Fig. 8. Application of star-delta theorem to a bridged-T network (a). The star of C and R is transformed to a delta (b) and gives the result (c). Alternatively, the delta of L,r,C can be transformed to a star as in (d).

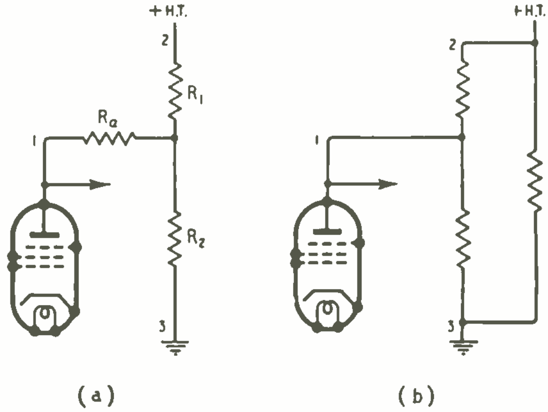


Fig. 9. The application of the star-delta theorem to the problem of Fig. 4. The star of (a) transforms to the delta of (b) and at once shows up the redundant resistance.

Fig. 10. Basic 4-terminal network.

this would call for $Z_{12} = 59 + 59 + 59^2/14.6 = 358\ \Omega$ and $Z_{13} = Z_{23} = 59 + 14.6 + (59 + 14.6)/59 = 88.2\ \Omega$.

The star-delta theorem is often of great use theoretically in simplifying things. This is especially the case in bridged-T networks. The circuit of Fig. 8(a) without the resistance R is a simple parallel resonant circuit which might be used as a rejector. The inductance L has losses which are represented by the series resistance r and it is tuned by the two capacitors C in series having the total value C/2. At resonance the circuit behaves as a high resistance of value $2L/Cr$, the dynamic resistance.

When R is added, it is possible to make it behave as though the dynamic resistance were infinite. Physically, some current passes through L and r and some through the T network C, R, C. By the adjustment of the components, the currents at the output can be made equal and opposite. The conditions are most easily determined by using the star-delta transformation, which can be applied in two different ways. The first is to transform the star of two capacitances and one resistance to a delta (b). We get

$$Z_{12} = 2/j\omega C - 1/\omega^2C^2R$$
$$Z_{13} = Z_{23} = 1/j\omega C + 2R$$

and can re-draw the circuit as Fig. 2(c). This is an exact equivalent of (a) but is not physically realizable in this form because it includes a negative resistance $-1/\omega^2C^2R$. From this, one can write down at once the conditions for resonance and infinite attenuation as

$$\omega^2LC/2 = 1$$

and $\quad r = 1/\omega^2C^2R$

or $\quad R = L/2Cr$

The alternative way of applying the transformation is to turn the delta of L, r and the capacitors into a star as in Fig. 8(d). In this particular case, this is not such a good transformation as the first, because the expressions for the star impedances turn out to be more complex than those for the delta impedances of (b).

Wherever such alternatives for transforming a circuit exist, it is quite usual for one to be simpler than another and one naturally chooses this one.

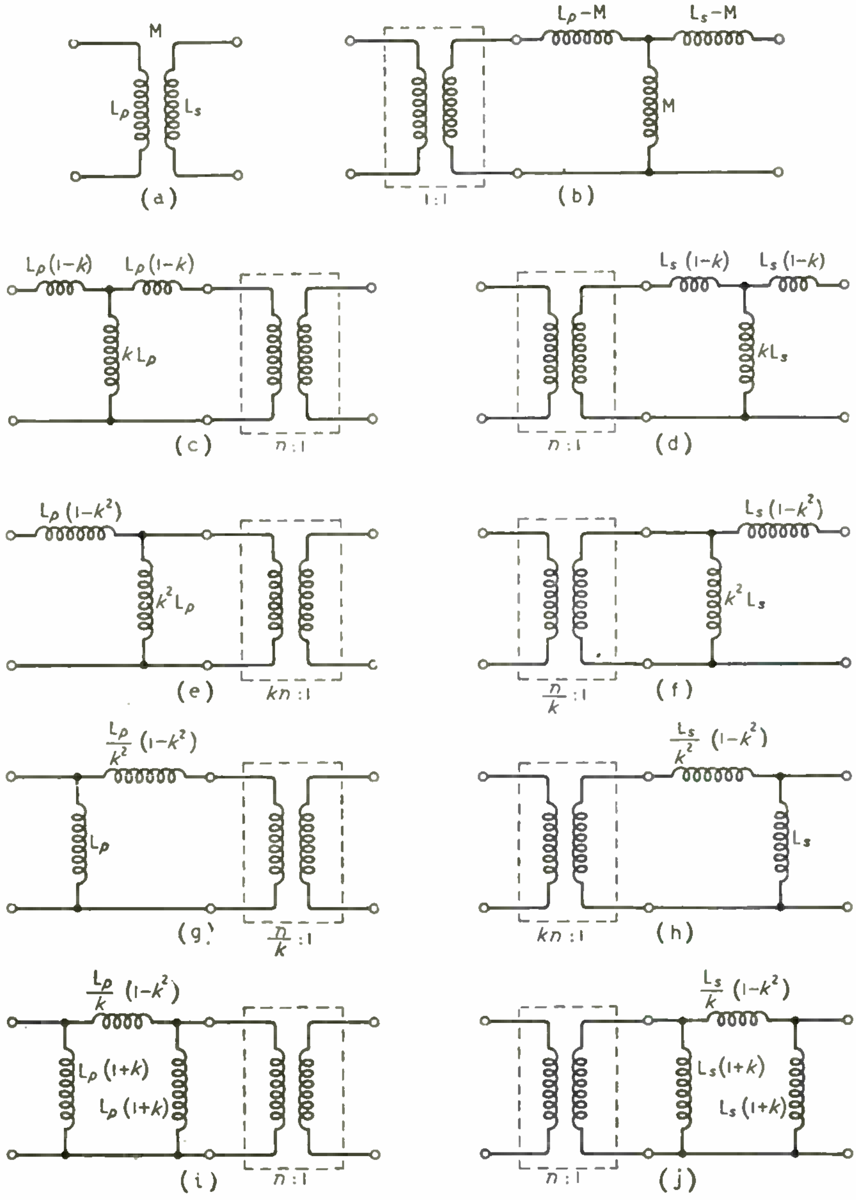
The star-delta theorem can be applied to the valve problem of Fig. 4 and, although it is not so useful as Thévenin's in this instance, it does show up more clearly the redundancy of one of the resistors in Fig. 4(b). The circuit is repeated in Fig. 9(a) and the three resistors form a star which can be replaced by a delta formation as in (b). It is at once obvious that one resistance R_{23} comes straight across the h.t. supply and performs no useful function. We can remove it, therefore, which is the same as making it infinite and the equations of Fig. 7 show that when Z_{23} is infinite Z_1 is zero. In Fig. 9(a), R_a is the Z_1 component and becomes zero. We end up then with the simple potential divider.

In this instance, the star-delta theorem is less useful than Thévenin's because it does not include the supply voltage and so does not permit us to calculate component values for particular conditions.

Equivalent Circuit Theorem

There is a theorem which states that any two circuits are equivalent if their open- and short-circuit impedances are the same. It applies to four-terminal networks, shown diagrammatically in Fig. 10. The open-circuit impedances are the impedances between 1 and 2 when 3 and 4 are open and between 3 and 4 when 1 and 2 are open. The short-circuit impedances are the impedances between 1 and 2 when 3 and 4 are shorted and between 3 and 4 when 1 and 2 are shorted. By the application of this theorem it is possible to prove a whole series of equivalent circuits for the transformer, some of which are of great utility. No less than ten such equivalents are shown in Fig. 11.

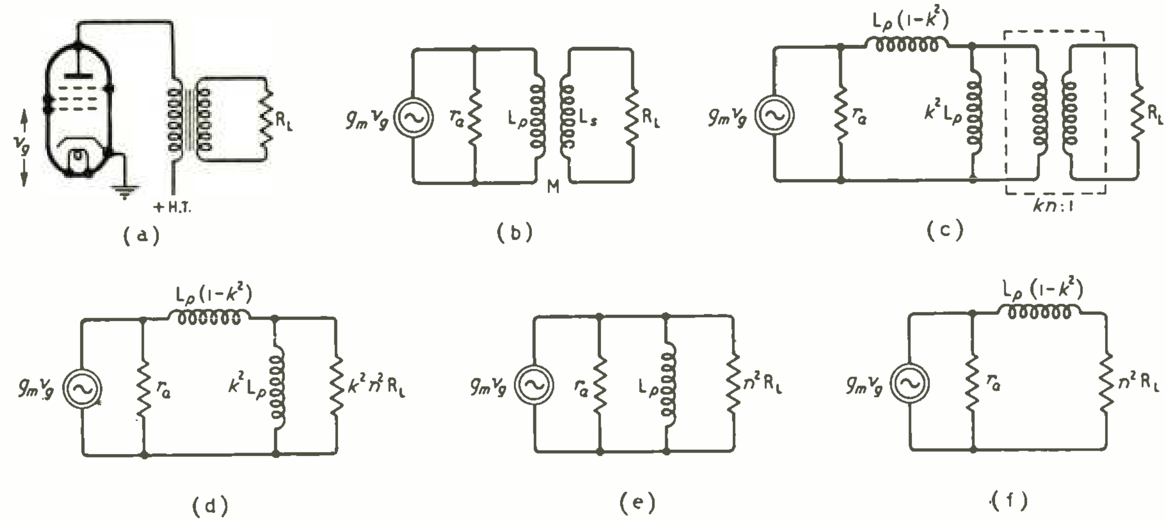
These circuits are all exact equivalents and the basic arrangement is shown in Fig. 11(a), a primary coil L_p and a secondary coil L_s having mutual inductance M between them. In the equivalents, the transformer shown in a dotted box is an ideal one which serves to preserve the d.c. isolation and to provide a voltage transformation ratio. It has no other



WHERE n = RATIO $\frac{\text{PRIMARY}}{\text{SECONDARY}}$ TURNS $= \sqrt{\frac{L_p}{L_s}}$

AND k = COUPLING COEFFICIENT $= \frac{M}{\sqrt{L_pL_s}}$

Fig. 11. Ten equivalent circuits for a transformer are shown here. The ideal transformer, surrounded by a dotted box, provides the turns ratio and d.c. isolation but has no other characteristics. The coupling coefficient is $k = M/\sqrt{(L_pL_s)}$ and the ratio of the primary/secondary turns is nominally n; actually $n = \sqrt{(L_p/L_s)}$.



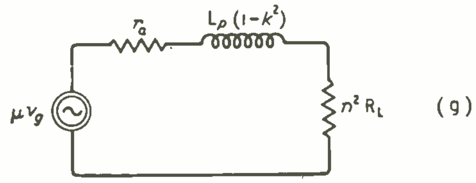


Fig. 12. Application of transformer equivalent circuit. At (a) a valve is transformer-coupled to a load R_L and the equivalent of this is shown at (b). The transformer part of this can then be changed to give the form shown at (c) and, in turn, this reduces to (d). For low and high frequencies respectively this can be further simplified to (e) and (f). When r_a is small, as with a triode, (g) is a more useful form than (f).

characteristics. The other inductances represent the other characteristics of the real transformer. In these equivalents k is the coupling efficient of the real transformer and equals $M/\sqrt{L_pL_s}$ while n is $\sqrt{L_p/L_s}$ and is usually the turns ratio of the real transformer. The true definition is $\sqrt{L_p/L_s}$, however, and there are cases where the two are not quite the same.

Some writers make great use of the symmetrical forms of circuit (c) and (d) but the others are often simpler and, in particular, (f) is a very convenient one.

These circuits are of considerable help when one wishes to determine transformer characteristics by measurement. From Fig. 5(f), for instance, it is obvious that if one measures inductance at the secondary terminals one measures L_s with the primary open and $L_s(1-k^2)$ with the primary shorted, and from the two k can be determined. From Fig. 5(e) similar measurements on the primary give L_p and $L_p(1-k^2)$, from which k can again be determined. Then, knowing L_p and L_s, n can be found.

One great use of these equivalent circuits is the way in which they make important factors almost obvious instead of being determinable only after a lengthy calculation. For example, suppose a valve is coupled to a load resistor R_L as in Fig. 12(a). We replace the valve by its Norton equivalent circuit, and the equivalent circuit becomes (b). We now use the transformer equivalent of Fig. 11(e) and get (c) and then transfer the load R_L from the secondary to the primary of the ideal transformer where it takes the value $k^2n^2R_L$, as in (d).

It is at once obvious that k^2L_p will rob $k^2n^2R_L$ of current, for it comes in shunt with it, while $L_p(1-k^2)$ will cause a voltage drop, since it comes in series. In practice, in a.f. applications k^2 is very nearly unity and $L_p(1-k^2)$ is consequently very small compared with k^2L_p. Because of this, the former has a negligible effect at low frequencies and the latter is negligible at high frequencies.

The circuit can thus be further simplified to the equivalents (e) and (f) of Fig. 12, which are valid respectively for low and high frequencies only. If r_a is large compared with n^2R_L it can usually be neglected. If it is not large, the Thévenin equivalent is better than the Norton at high frequencies and (f) can be changed to (g).

It is plain from (e) that the frequency response at low frequencies depends only on the relation of L_p to the value of r_a and n^2R_L in parallel and that at the lowest frequency $f(=\omega/2\pi)$ we must have ωL_p large enough compared with this resistance for it to shunt it negligibly. It is equally clear from (g) that the leakage reactance at the highest frequency must be small compared with $r_a + n^2R_L$; if it is not, it will reduce the current.

If the lowest and highest frequencies are fixed, L_p must be fixed by low-frequency requirements and $L_p(1-k^2)$ by high frequency, which means that k is fixed by the high-frequency needs. More strictly, k is fixed by the bandwidth required.

The use of the transformations brings out the important factors very simply and clearly without the use of appreciable mathematics. From (e) and (g) it is easily possible to write simple equations expressing the performance from which numerical values can be obtained.

LETTERS TO THE EDITOR

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

Output Stage Performance

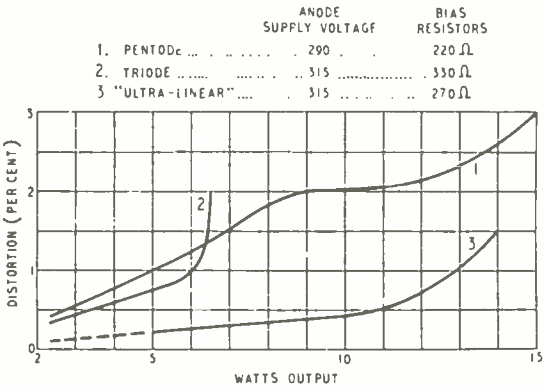
IN a description of the Osram 912 amplifier given in the September issue (page 430), it is stated that the output circuit is a compromise between pentode and triode operation as far as efficiency and harmonic distortion are concerned.

The object of employing the "ultra-linear" circuit in the Osram 912 is to provide virtually the same efficiency and power as pentode-connected valves, but a distortion lower than that of either pentodes or triodes at their respective full outputs.

Some data relevant to the Osram N709 valves specified for this circuit are as follows:

	Pentode	"Ultra-linear"	Triode
Anode input, per pair	33.4	31.5	25.2 watts
Output power	15	14	6.5 watts
Distortion, full output ..	3	1.5	2 per cent
Distortion 6.5 watts approx.	1.5	0.25	2 per cent
Output impedance	38,000	9,000	4,000 ohms
Grid-to-grid input voltage	14	18	18 (r.m.s.)

The accompanying curves enable a comparison to be made at other power levels. These results are for the output stage alone, without external feedback.



Other valves, such as the KT66, also work well in the circuit.

The output transformer is connected to the anodes and screens and not to the cathodes as stated.

GRAHAM WOODVILLE.

The M.O. Valve Co., Ltd.

"*Inexpensive 10-Watt Amplifier*"

WITH reference to the second part of E. F. Good's letter in your October issue concerning "the obstinate refusal to adopt the tertiary feedback-winding system" as used in P. J. Baxendall's amplifier, I would suggest the reason why this ampifier did not become as famous as others such as the Williamson is not connected in any way with the method of obtaining feedback, but because of the use of beam-tetrodes in the output stage.

With the tetrode output and about 40 db negative feedback the Baxendall amplifier has roughly the same distortion as triode output types, such as the Williamson, *on a resistive load.*

With a variable impedance device such as a loudspeaker, the load presented to the output valves rises to a high value at the high audio frequencies, and also at the bass resonant frequency. With a triode-output stage and to a lesser extent with the triode-tetrode connected type, any increase in load above the nominal causes a reduction in distortion; but with the tetrode or pentode a violent increase in third harmonic results. Negative feedback only reduces distortion by a given factor (×100 for 40 db) and does not eliminate the cause.

Mr. Baxendall states that when his amplifier was tested on a loudspeaker load, "a several times increase in distortion of the output voltage occurred due to the non-linearity distortion in the current drawn by the loudspeaker" (*W.W.*, Jan., '48). I am not so sure that all the blame for this can be attributed to the speaker.

I will admit that the RC network across the primary of the output transformer in the Baxendall amplifier keeps the load reasonably constant with increase of frequency, but this does not alter the situation at the l.f. end. I am also surprised that the new Mullard amplifier circuit contains no such load correctors.

From the above reasoning it seems that to get equivalent performance to a triode from a tetrode or pentode output stage on loudspeaker load, considerably more feedback is required than is apparently necessary from comparative tests on resistive loads. This, of course, is not practicable with a single-loop feedback circuit. A multiple-loop feedback arrangement, including one loop in the output stage, by triode-tetrode connection, seems to be the best answer to the problems of size, weight, and heat developed.

A further advantage of the multiple feedback loop system is that those unpleasant peaks outside the audio band can be avoided. The rise in response below 10 c/s with the Williamson circuit is the main reason why the complication of an additional mains transformer and smoothing system is necessary for the h.t. supply to auxiliary apparatus.

Aldeburgh, Suffolk. JOHN BRIGHTON.

"*Why Lines?*"

AS one who has long been interested in Lissajous scanning and who recently has spent considerable time observing television pictures produced in this way, you may imagine my surprise at seeing this article in your August issue. Why lines, indeed?

As I read further through the author's list of supposed advantages, and the Editor's list of probable difficulties, I noted that he, too (as I) had failed to anticipate a major objection to the scheme. Here is a short account of the work which brought this to light.

The possibility of producing television signals with essentially no geometrical distortion is very attractive for various industrial applications. The simplicity of the equipment (which appealed so much to Mr. Hughes) had made the idea appear very promising. Hence we constructed four sine-wave amplifiers to drive the horizontal and vertical deflection coils of a flying-spot scanner and monitor kinescope and used separate audio oscillators, set at approximately 11,350 c/s and 15,250 c/s. It at once became evident that the phasing was very critical, in order to avoid a double image due to mis-register between the forward and reverse scans. It also turned out that the frequencies must be held very nearly correct to prevent the pattern from degenerating into a lower-order coarse pattern or, worse still, a very badly flickering one, if reasonable brightness were used. We then built a synchronizing generator to produce locked signals, and while this never did work perfectly, we were nevertheless convinced that no insuperable obstacles stood in the way of a satisfactory solution of the scanning problem.

All observers remarked favourably on the geometry of the picture and noted with genuine surprise the fact that there was no need to make any brightness correction due to non-linear scanning. Even on a blank raster (no video signal) the pattern appeared uniformly bright except at the very edge, and when a picture was present even this effect was practically invisible.

You might reasonably ask, in view of the above favourable results, what is wrong with the system? It is this: Each point in the picture is scanned twice. At the centre, for example, the first scansion occurs as the beam is moving down to the right and the second as the beam moves up to the right. Nearby points may be scanned to the left. Now if the picture detail at this point happens to be a straight narrow line, those segments reproduced by the first scansion are shifted slightly down to the right, and alternate segments reproduced by the second scansion are shifted up to the right, due to the finite bandwidth of the system. The line is reproduced as a zig-zag, and the entire scheme is shifted into uselessness. This effect is most noticeable on a test chart, such as the RMA Resolution Chart, 1946. On low-detail subjects such as close-up portraits, the effect is not serious and merely gives a "soft" rendering.

In passing, it is worth noting that sinusoidal scanning, rather than being entirely novel, was analyzed in the report of the first N.T.S.C., "Television Standards and Practice," McGraw Hill, 1943, p. 33, and was rejected at that time for reasons similar to the Editor's list of probable difficulties.

W. F. SCHREIBER.

Technicolor Motion Picture Corp.,
Hollywood, U.S.A.

"*Filters Without Fears*"

I HAVE read with great interest the article by Thomas Roddam in your September issue, in which the application of Chebyshev polynomials to filter theory is clearly and explicitly given. The purpose of my letter is to point out that the above spelling of this distinguished mathematician's name has nothing to do either with the "post-revolutionary alphabet" or with "foreign politics." A slight acquaintance with the Cyrillic alphabet will show that only minor changes were made at the time of the revolution, none affecting the name in question, and all made with the laudable object of removing redundant letters. Transliteration from non-Roman alphabets is always a matter of some difficulty and should aim at rendering, letter by letter, words printed in one alphabet into another. It is desirable in this process to obtain a result which will also enable the reader of the transliterated material to pronounce it as nearly as possible in the original way. It is unfortunate that many of the Russian proper names with which we are familiar came into Western literature *via* German transliteration, Tchebycheff being a case in point. Librarians and linguists in English-speaking countries have devised consistent systems of transliteration based upon English phonetics and conforming to the above criteria. These are widely used—as reference to *Science Abstracts* will confirm—and the spelling Chebyshev is the recognized English transliteration, which should be universally used in place of the German monstrosity given by Mr. Roddam.

The further point that the polynomials are denoted by T_n (a) is irrelevant. After all, the Bessel functions are J(x) and Y(x), in no way suggesting their discoverer; and the elliptic functions have a notation far removed from the names of Jacobi or Weierstrass.

Glasgow University. B. HAGUE.

Television Interference

G. O. THACKER'S letter in the October issue of the *Wireless World* concerning the radiation of interference from television receivers raises a further point.

If it were possible to comply with Condition 4 of the broadcast television licence, which states that "the apparatus shall be so maintained and used that it does not cause interference . . ." the television pirate detector vans described and depicted on p. 476 of the same issue would need to be rather more complicated and probably less effective.

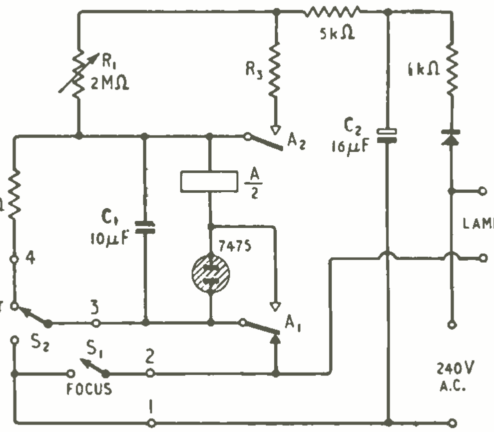
I wonder if this is why the G.P.O. has limited powers in preventing television generated interference?

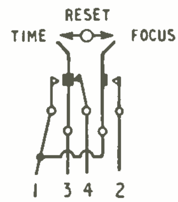
Further, it would seem ironical that, as the law stands at the moment, following the detection of a pirate television receiver by the existing means, a summons could be issued on two counts (a) the use of a television receiver without a licence, and (b) failing to comply with Condition No. 4 governing the use of a television receiver.

Stoke-on-Trent, Staffs. G. E. KING.

"*Neon Timers*"

WITH reference to the circuit in Fig. 3 of the article by B. T. Gilling (your September issue, p. 460), this could be made much simpler and still perform the same





Circuit of the simplified photographic timer. The key on the left could be used in place of S_1 and S_2 and would be connected to the numbered points on the circuit.

functions by using the arrangement shown in the accompanying diagram.

Resistor R_3 would depend on the resistance of relay A, and is adjusted to allow a current greater than its holding current to flow to hold the relay after C_1 has discharged. The resistance of the relay is not critical. The voltage regulator has been dispensed with as any inaccuracies in the time lag would be due to the firing potential of the 7475 tube and the c.r. time constant of C_1R_1. Also, I do not think the mains supply voltage varies sufficiently to justify its use for the order of accuracy needed for photographic purposes.

The operation is as follows. When S_2 is thrown to the TIME position, C_1 charges via R_1 until the 7475 tube fires, when the relay operates and locks via S_2, A_1, the relay coil, A_2 and R_3. On throwing S_2 to the RESET position, the relay releases and C_1 discharges via R_2. S_1 is used leaving S_2 in the RESET position for focusing without timing. S_2 could be a change-over toggle switch and S_1 a make-break switch, or a single-lever key could be used.

Norwich. J. R. BARNARD.

Flywheel Synchronizing

2.—*Principles of Automatic Frequency Control*

By W. T. COCKING, M.I.E.E.

In practice, flywheel synchronizing is normally obtained through the use of an automatic frequency control (a.f.c.) system. In this, the sync pulses are compared in a phase discriminator with a locally-generated voltage which is usually obtained from the line timebase. As a result of the comparison, an error signal is developed which depends upon any difference of frequency or phase between the sync and the local waveforms. This error signal is passed through a filter, which provides the flywheel effect, and is then applied to the timebase as a control voltage which operates to bring any frequency error to zero.

The system is not, however, the same as the one which is used to control the frequency of the local oscillator of some superheterodyne receivers, in spite of the fact that that is also called an automatic frequency control system. In that, a control voltage is developed which depends upon the difference of the frequency generated from its proper value and the circuit cannot reduce an error to zero, for some error must exist for there to be a control voltage at all. This kind of circuit acts only to reduce the magnitude of an error but it cannot bring it to zero.

In the case of a timebase, it is essential that the frequency be exactly that of the sync pulse recurrence. The frequency error must be zero. Frequency and phase change together and it is not possible for one to change without the other changing also. It is, however, possible to have two frequencies which are exactly the same but which have any desired constant-phase relation. It is the relative phase of the local waveform and the sync pulses which is used, therefore, in order to develop a control voltage. Because of this, the system is sometimes known as one of automatic phase control, although it is not phase which is controlled, it is phase which is the controlling quantity.

In general form, all a.f.c. systems can be represented by the block diagram of Fig. 1, but sometimes an amplifier is included between the low-pass filter and the timebase. It can be seen from this that the circuit includes a closed loop, for the output of the phase detector depends on both its inputs and one of them is derived from the timebase which is under the control of the output via the low-pass filter. It is, therefore, a feedback system and, as in all such systems, the problem of obtaining stability is important.

There are many forms of a.f.c. circuit and it is not practicable to discuss all of them here. So far as possible the discussion will be in general terms and we shall endeavour to find the most suitable type before coming to the particular.

For correct synchronism, the start of flyback in the timebase must occur nearly in coincidence with the leading edge of the sync pulse. It need not coincide exactly but, in general, it must not occur more than 1 μsec earlier nor more than 4 μsec later. Any difference greater than this is likely to result in a noticeable displacement of the picture on the raster and, possibly, to fold over or cut off. The total difference of timing can thus vary over a range of about 5 μsec at most. When the two frequencies are the same, the phase detector produces an output which is dependent on the relative timing (that is, the relative phase) of two particular parts of the waveforms. This is illustrated in Fig. 2, in which the sync pulse waveform is shown at (a) and the saw-tooth wave of the timebase in (b), (c) and (d). In (b) the relative timing, or phase, is such that the start of flyback coincides with the leading edge of the sync pulse, which is the normal condition with direct locking. In (c) the flyback is shown starting about 1 μsec earlier and in (d) about 4 μsec later.

Let us now consider a timebase which has no synchronizing system at all, but a manual hold control which is capable of very fine adjustment. As this control is turned, slowly, the frequency of the timebase comes nearer and nearer to the correct value and at length equals it. The timebase is then in synchronism with the signal and the frequency error is zero.

The saw-tooth and the sync signals are not necessarily in their right relative phase, however, and the sync pulses may well appear somewhere remote from the flybacks. The picture will then appear divided into two parts separated by a vertical black bar which corresponds to the blacker than black of the sync pulses and the black of the back and front porches. What should be the left-hand part of the picture will appear to the right of the bar and what should be the right-hand part will be on the left. The picture will be quite coherent, however.

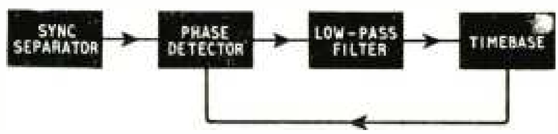


Fig. 1. Basic form of an a.f.c. system for flywheel sync.

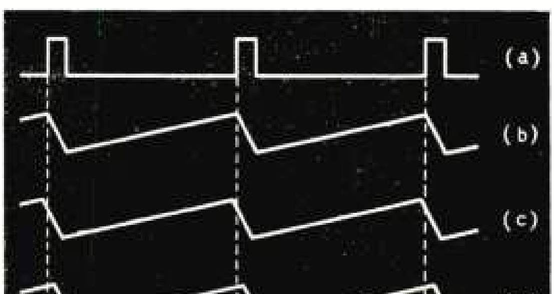


Fig. 2 Sync pulses are shown at (a) with a saw-tooth in correct phase relation at (b). The limits of permissible phase error are indicated in (c) and (d).

To obtain the correct picture it is necessary to bring the sync pulses and flybacks into approximate phase coincidence, as in Fig. 2. In order to do this, the timebase frequency must be slightly altered by the hold control and then brought back again to its correct value by returning the control to its previous setting. Synchronism is then established again but with a different phase condition.

The attainment of the correct frequency demands the provision, by the hold control, of a certain steady voltage, the precise value of which depends on many factors, such as supply voltage, temperature, stability of components, etc. If any of these change the "steady" voltage must change in a compensating manner.

Correct phasing can be secured only by a momentary change of frequency. If the black bar occurs in the centre of the picture, for instance, it is theoretically possible to bring it to its proper place at the edge by making one individual scanning line of one-half its normal duration. The change of "frequency" can occur and be over all in less than one line, which implies a momentary change to double the normal value. Alternatively, the change can be minute and persist for many lines, so that the black bar gradually creeps to the edge.

In practice it is not easy to observe these effects, but they can be seen if the timebase is a very stable one and the hold control permits sufficiently fine adjustment to be made.

An a.f.c. system performs the equivalent action to adjusting a manual control by providing automatically the voltage which would otherwise be provided by the manual control. When there is no frequency error, a voltage is developed by the discriminator which has a constant value depending only on the phase difference between the sync pulses and the saw-tooth waveforms. The system then settles down in synchronism, but with a phase error which depends upon the voltage needed by the timebase for it to run at the correct frequency. If anything changes to make the timebase need a different voltage, the relative phase must change so that this new voltage can be provided. While the change is actually occurring there is, of course, a change of frequency also.

Noise Reduction

It is necessary that synchronism should be obtainable only when the flyback is in approximate coincidence with a sync pulse and that the range of possible phase errors should be small. This is easily arranged by using the sync pulses on the one hand and a waveform derived from the flyback on the other as the signals on the phase discriminator, for a steady output can then be obtained only when at least some parts of them occur together.

If the manual hold control is adjusted while the a.f.c. system is in operation it is found that the visible picture moves as a whole sideways on the raster by a small amount. This is a characteristic of all a.f.c. systems and results from the change of phase needed to maintain the frequency at its correct value.

At this stage it will probably not be at all evident how an a.f.c. system reduces the effects of noise and interference. It does so in reality in substantially the same way as with the tuned-circuit system of Part 1. The sync pulses, together with the noise and interference, are mixed in some way with the local waveform in the phase discriminator and the output is usually some form of pulse having some characteristic dependent on the relative phase of the signals, but it is also accompanied by noise and interference. The control voltage for the timebase is the mean value of the discriminator output and it is virtually d.c. It is the filter which changes the pulse output to d.c. and it does this by smoothing the wave. On a frequency basis it attenuates all the varying components to leave the d.c. and, in doing so, it naturally removes the noise and interference.

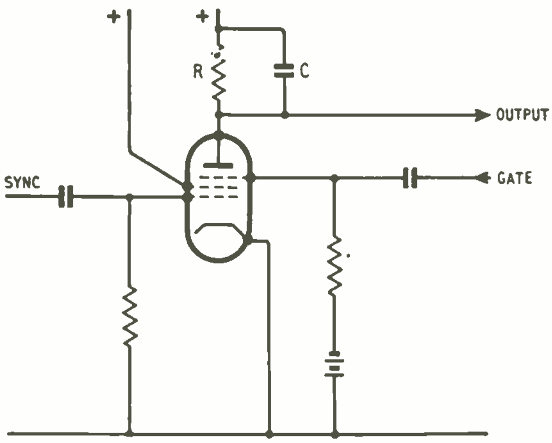


Fig. 3. Simple form of phase detector. Sync and gating pulses are applied to the control and screen grids of a pentode.

An analogy is helpful here. In a sound broadcast receiver a heterodyne whistle results if two stations are too near together in frequency. This whistle can be eliminated by making the r.f. circuits so selective that one r.f. carrier is sufficiently attenuated relative to the other, or it can be removed by using an a.f. filter after the detector. The first is analogous to the tuned-circuit flywheel system described in Part 1 and the second to the a.f.c. type of flywheel circuit. Both operate by frequency selectivity, but in different places. It is much easier to obtain the necessary selectivity in an a.f.c. system, however, and a simple RC network suffices.

There are a great many different a.f.c. circuits and the differences are chiefly in the form of the phase discriminator. The two most important divisions are between balanced and unbalanced discriminators; within these, most of the different kinds fall into the unbalanced category but the balanced type is probably the more widely used.

It will appear later that the balanced discriminator has important advantages over the unbalanced and we shall deal mainly with this kind. It is not, however, the simplest to understand and it is helpful first to consider a particular form of unbalanced circuit.

The basic circuit of this is shown in Fig. 3. The sync pulses are applied to the control grid of a pentode; they are not differentiated and so the pulses are substantially rectangular. The pulses are negative-going and the cathode current is cut off whenever a pulse is present, but at all other times the grid is at about cathode potential.

The timebase waveform comprises positive-going pulses which are preferably, but not essentially, of rectangular form. This gating-pulse waveform is applied to the suppressor grid of the pentode which is so biased that anode current is cut off except when a pulse exists on the suppressor.

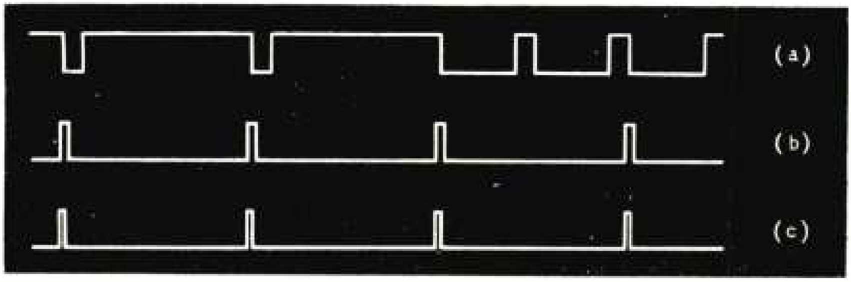


Fig. 4. Waveforms for the circuit of Fig. 3. are shown here. The sync pulses are at (a) and the gating pulses at (b). The anode-current waveform of the phase detector is shown at (c); the pulse width depends on the overlap of (a) and (b).

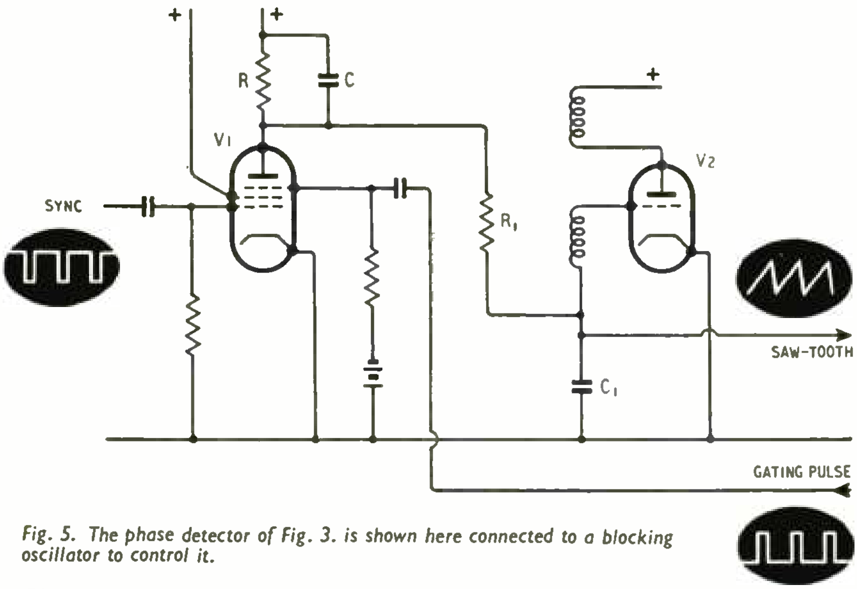


Fig. 5. The phase detector of Fig. 3. is shown here connected to a blocking oscillator to control it.

The anode *current* waveform is shown at (c) and comprises pulses of 2-μsec duration. Current flows for the first 2 μsec of the gate pulse because the sync pulse has not then started; the control grid is at about cathode potential and the current starts when the gate pulse brings the suppressor grid to cathode potential. No current flows for the last 2 μsec of the gating pulse for, after the first 2 μsec, the sync pulse comes along and cuts off the current.

No change of current waveform occurs during the frame-pulse period, for the gating is accomplished on the leading edges of the pulses and these recur regularly during the frame-pulse period. The half-line pulses have no effect because there is no gate pulse when they occur. Generally, noise or interference can have no effect at all except when it occurs during the time for which the gate pulse exists. This alone can result in a considerable reduction in the frequency with which noise and interference can affect the synchronizing, although it does not alter the magnitude of its effect when it actually does occur.

If a phase change occurs between the sync pulses and the gating waveform, the effect is of a relative displacement of (a) and (b) in Fig. 3. If (b) is moved to the left, for instance, the pulse overlap is reduced and the current pulses are longer. If it is moved to the right, the overlap is greater and the current pulses are shorter. The maximum permissible movement is the width of the gating pulse—4 μsec in this instance. It can move to 2 μsec later than the position shown, when anode current is just cut off completely, or to 2 μsec earlier when the anode current flows for twice the time shown in (c) and so has twice the *mean* value.

The resistor R in Fig. 3 might have a value of 100 kΩ and the peak anode of this valve might be 5 mA. The mean voltage across R for 100-μsec line period would be $500 \times 2/100 = 10$ V for the mean condition and would vary from 0 to 20 V as the phase varied from 2 μsec late to 2 μsec early.

If the timebase is normally running so that condition (c) is obtained and some change occurs, so that the gate pulses start arriving later, it means that the interval between them is increasing and the timebase is tending to run at a lower frequency. The resulting decrease of voltage across R, which is a rise of anode potential with respect to the positive h.t line, must be applied to the timebase so as to increase its frequency.

One way of doing this is to return the charging resistor of the timebase to the anode of the valve in

If sync pulses are absent, the valve passes anode current for the duration of every gating pulse and a mean voltage drop appears across R corresponding to the mean current through it. When sync pulses are present and overlap the gating pulses anode current is cut off during the sync pulses and so current flows only during that part of a gating pulse which is not overlapped by a sync pulse. Since the pulses are all of constant amplitude the magnitude of the anode current is always the same whenever it flows. The time for which it flows, however, depends on the overlap of the pulses and therefore, also, the quantity of electricity passed by the valve and carried into C. The resulting mean voltage developed across RC, which forms the output voltage, therefore, depends upon the degree of overlap of the pulses; that is, upon the relative phase of the sync and time-base waveforms.

Phase Discriminator Operation

One condition of operation is shown in Fig. 4. The sync pulses are shown in (a) with some of the frame pulses as well as the line. The gate-pulse waveform is shown at (b); it is generated by the timebase and is here rectangular and the pulses are assumed to have a duration of 4 μsec, compared with the 10 μsec of the line sync pulses. The phase relation shown is with the gate pulses centred on the leading edges of the line sync pulses.

Fig. 3. The increase of voltage then increases the charging current and the capacitor of this circuit charges more rapidly. This means, of course, that the frequency of the timebase tends to rise.

If the initial charge is the other way round, everything happens inversely.

The interconnection of the circuit of Fig. 3 with a blocking oscillator saw-tooth generator is shown in Fig. 5, in which R is the charging resistor and would normally be in part variable as a manual hold control. The circuit is not a very practical one for it is difficult to generate a suitable gating pulse.

Normally with direct locking the flyback is initiated by the leading edge of the sync pulse and may start almost immediately or after a small delay. The picture ceases 0.5-1 μsec before the leading edge. It is usually impracticable to make the leading edge of the gate pulse occur prior to the start of flyback and it is not easy to make it even coincide with it. The easy thing is usually for it to occur several microseconds after the start of the flyback.

When the leading edge of the sync pulse is gated and the start of the gating pulse coincides with the start of the flyback, then the flyback must start earlier than the sync pulse by up to the duration of the gating pulse. In the example quoted, flyback must start from 0 to 4 μsec before the sync pulse. As a result, from 3-3.5 μsec of the right-hand side of the visible part of the picture may be cut off or folded over. This can be avoided only by narrowing the gating pulse and it should not exceed 1 μsec in duration and, preferably, be still less. The output of the phase detector is proportionally reduced and will quite likely not be enough for adequate control.

It would, of course, be possible to obtain the same result by delaying the sync pulses and this would be preferable if it could be done easily enough. At present, however, it seems to be economically impracticable.

Another difficulty arises over the generation of a 1-μsec undelayed pulse by the timebase. The gating waveform is usually taken from the output stage of the timebase, for a large voltage exists there and it is buffered from the saw-tooth generator by the output stage. The pulse has considerable delay over the start of the flyback, however. If a voltage from the saw-tooth oscillator itself is used, there is a grave risk of the oscillator being triggered by the pulses in the phase detector, for the smoothing by RC of Fig. 2 is often far from perfect and appreciable amplitude of pulse does exist here. The circuit would then merely degenerate to an expensive and unsatisfactory way of obtaining direct locking. Because of these practical difficulties the circuit is not much used.

Effect of Noise

It will be remembered from Part 1 that the effect of noise and interference is partially or completely to fill a sync pulse or, by cancelling the signal just prior to a sync pulse, to make the pulse in effect start earlier than it should do. In the gating circuit the effect is, therefore, to make the anode current pulses somewhat variable in width. Noise and interference do affect the mean output voltage and hence the operation of the time base, but not to the same degree. Because of the integrating effect of CR, Fig. 2, any effect is very slow and spread over many scanning lines. The effect can be, in fact, no more than a slow and small sideways movement of the picture as a whole. If the time constant is smaller, so that the integration is over a period corresponding to a dozen lines or so only, then there may be a displacement of these lines relative to the others, but it is not an abrupt and erratic displacement like that which occurs with direct locking. It is a gradual displacement and the visible effect may be more than the appearance of a slight bend in vertical lines.

If, on the average, the effect of interference is to cause sync pulses to be as much early as they are late, then because CR gives an averaging effect, the effect of the interference is zero. In practice this is not likely to occur, but some reduction in its effects may well occur because of the two different ways in which noise and interference can affect the pulses.

(To be concluded)

Headquarters for Scottish Electronics

ON October 11 the Duke of Edinburgh opened a new laboratory building adjacent to the Ferranti works at Crewe Toll, Edinburgh. It has been built by the Government, and as well as housing Ferranti's own research staff it will provide a centre for the Scottish Council scheme of developing the electronics industry in Scotland. As reported in our February, 1952, issue, this scheme is based on the fact that the Scottish Council have induced the Government to place a fair share of research contracts in Scotland, from which firms will be able to build up their technical knowledge and facilities. Once the firms have established themselves with this kind of work they will be in a position to go into the commercial applications of electronics.

Ferranti's are acting as a "parent" organization for the scheme and their main function is to accept large design contracts from the Government and sub-divide them amongst the firms participating. They also provide technical liaison in the progress of these contracts and assist the participating firms with administrative matters.

The importance of the new laboratory block to the scheme is that it will provide a place where engineers from these firms can work alongside Ferranti engineers and so gain experience of electronic techniques which they would not otherwise obtain. The scheme is, in fact, mainly intended for existing firms who wish to establish their own electronics departments. One engineering firm, for example, which started by appointing one man to work in the Ferranti laboratories, now has a team of about 30, including six graduates, at work in its own factory.

Sunspot Minimum

SUNSPOT activity has, on the average, been decreasing since the maximum in May, 1947, and by the beginning of this year had reached a very low level.

Although it is not possible to predict future sunspot activity with any certainty it had been generally expected that it would reach a minimum value sometime during 1954. According to information received from the Zurich Observatory the activity is now increasing again, and it is thought probable that the minimum occurred in June. It is too early yet to be certain about this, and it is also improbable that sunspot activity will increase sufficiently to have any very significant effect upon the usable radio frequencies for several months to come.

The significance of sunspot activity to radio men is, of course, that it is one of the observable phenomena indicative of the general solar activity, upon which depends the degree of ionization of the reflecting layers of the ionosphere. When the solar activity is high the ionization of the layers is greatest, and the higher frequencies become usable for long-distance communication, when solar activity is low the m.u.fs. are lower. T. W. B.

Filters Without Fears

3.—*Some Practical Design Calculations*

By THOMAS RODDAM

IN the two previous parts of this article which appeared in the August and September issues, some attempt has been made to show how low-pass filter networks can be designed from first principles. The first article, which contained two stupid slips at the beginning, was intended to lay a foundation for its successors and in this article the idea of the Butterworth, maximal flatness, response was introduced. The second article dealt with the more complicated Tchebycheff response, the closest approximation type of characteristic. The *Wireless World* " standard reader," who serves much the same purpose as the canary in the coal mine or the leech in the jam jar, has been heard muttering that all this mathematics would never have done for *The Signal*. Can I introduce a new non-symbolic algebra, to serve as a counterpart to that study now so popular among electronic engineers, symbolic logic? Dare I press on with the mathematics and tell those of you who find it too heavy going that I can do " nothing but sympathize?"

Let us at least see where we have reached. The basic circuit which we are considering is made up of a generator of impedance R_1 and a load, R_2. Across the load there appears a voltage V_2, produced by the generator voltage V_1. Quite obviously,

$$(V_1/V_2)' = (R_1 + R_2)/R_2$$

(So obviously indeed that I wrote it incorrectly in Part I).

This is the scene before we put in the filter, which consists of a chain of shunt capacitances and series inductances making a total of *n* reactive elements. which we call an *n*th order filter. For this filter we find a new expression (V_1/V_2) = some expression containing the frequency. The insertion loss of the filter is defined as

$$20 \log (V_1/V_2) - 20 \log (V_1/V_2)' \text{ decibels.}$$

If we were not being too strict about the exact meaning of the decibel, we could say that this meant simply the loss from generator to load with filter minus the corresponding loss without filter. Just at this time, however, the definition of the decibel is under scrutiny, so we must be careful with our words. For algebraic convenience, we can rewrite the expression above as

$20 \log [(V_1/V_2)/(V_1/V_2)']$ (In Part I, I'm afraid, the prime was in the wrong place).

Now the expression $(V_1/V_2)/(V_1/V_2)'$, which we call N for short, is a complex quantity which contains terms in j, j^2, and so on. Of course $j^2 = -1$ and $j^3 = j \cdot j^2 = -j$ but after getting rid of the j^2's we finish up with $N = A + jB$, where A and B are expressions containing ω. The physical meaning of this is that N contains information about both the insertion loss and the insertion phase shift, and it is all mixed up together. We want to know the insertion loss, so we take $|N| \angle \theta = \sqrt{A^2 + B^2}$ arc tan B/A and the insertion loss is $20 \log |N| = 10 \log |N|^2 = 10 \log (A^2 + B^2)$. If for any reason we want to work with the insertion phase shift, we have $\tan \theta = B/A$.

For the second-order filter we derived the two equations

$$V_1/V_2 = (R_1 + R_2)/R_2 + j\omega(CR_1 + L/R_2) - \omega^2 LC$$
$$(V_1/V_2)' = (R_1 + R_2)/R_2$$

and dividing one by the other,

$$N = 1 + j\omega[CR_1R_2/(R_1+R_2) + L/(R_1+R_2)] - \omega^2 LC\, R_2/(R_1 + R_2)$$

Therefore $|N|^2 = 1 + \alpha\omega^2 + \beta\omega^4$, where the actual expressions for α and β were given in Part I. By putting $\alpha = 0$, the expression for $|N|^2$ is simplified to $1 + \beta\omega^4$, which is what we call a Butterworth response. The equation $\alpha = 0$ fixes a relationship between the inductance L and the capacitance C, and this relationship was given at the beginning of page 369 (August issue). It depends on the ratio of the two resistances R_1 and R_2, and some special cases were considered.

It is fairly certain that $\alpha = 0$ leads to the simplest solution, but is it the best? Bitter experience suggests that because it is the simplest it will not be the best. A more complicated solution is obtained by using the Tchebycheff polynomials, which are expressions giving a response oscillating between limits in the pass band, and offering us a choice of those limits. In Part II I used the limits of ± 0.625 db mainly in order to make the mathematics simple. This gave us two equations, $\alpha = -1$ and $\beta = +1$ and these in turn led to a new relationship between the inductance and the capacitance, a relationship which depends on the ratio of the two resistances.

Butterworth or Tchebycheff?

Have we wasted our time with the extra complication, and was the whole exercise worth while, anyway? The results which were derived in Part II showed us that the Tchebycheff filter has a much narrower transition region than the Butterworth filter. We can perform a different sort of calculation to compare the two types of filter and to give an idea of the sort of problem we may want to tackle. Suppose that we are dealing with the input to a video amplifier and we have a source with an impedance of 1,000 ohms. which is to be connected to the input grid, with a capacitance of 10 pF. In this particular example matching will be regarded as unimportant: what we shall look for is bandwidth. Common sense tells us that we should not put any resistance into the circuit, because that will reduce the available signal, so that in all the results derived in the previous articles we can take $R_2 = \infty$.

Let us consider what we will get. The simplest solution is just to connect the source directly to the input grid, giving us a first-order filter. It will be 3 db down when $2\pi f CR_1 = 1$ or $f = 1.6$ Mc/s. What is more, and more disturbing, it gives us a very gentle cut-off, and the response will be 1.25 db down when $10 \log [1 + (2\pi f CR_1)^2] = 1.25$ or $f = 0.92$ Mc/s.

We can do better than this by adding a series inductance to convert the network into a second-order filter. With a Butterworth filter we find that the response is 3 db down at $2\pi f CR_1 = \sqrt{2}$ or $f = 2.26$ Mc/s. It is a rather sharper cut-off, too, and at the 1.25 db. point we have $f = 1.72$ Mc/s. Another capacitance, across the input end of the network, would give us a third-order filter, with $f_{3db} = 2.4$ Mc/s and $f_{1\cdot25db} = 1.95$ Mc/s.

For the Tchebycheff response, we have a tolerance of ± 0.625 db, and this gives us a response of the form $1 - x^2 + x^4$ with a 1.25-db bump at $x = 0.7$. The edge of the pass band is at $2\pi f CR_1 = 1$ or 1.6 Mc/s.

For this tolerance, therefore, Butterworth seems to give a higher limiting frequency than Tchebycheff. Against this we must set the fact that the Tchebycheff response is always above the zero frequency level, while the Butterworth response is below it: the reduction in bandwidth is paid for by a small increase in gain.

When we consider very much closer tolerances on frequency response, in which there is no gain difference worth worrying about, we find that the Tchebycheff responses do offer some advantage in terms of bandwidth for a given capacitance-resistance situation. The arithmetic is complicated, and I don't propose to do more than assure you that it is so, and that facts and figures are given in "Amplitude-Frequency Characteristics of Ladder Networks" by E. Green, published by Marconi's Wireless Telegraph Company.

Tolerance Calculations

We can use all this algebra for interstage coupling networks, too, if we stick to third-order filters. One reader wrote to me and complained that R_1 was never really infinite, even with a pentode, and of course R_2 cannot really be infinite, what with grid leaks, transit-time damping, dirty valve bases and anything else you choose to mention. The point which the reader made was that you need to get the volts on to the anode, so you must put in a physical resistor. Partly, of course, it all turns on tolerances: the tolerance on valve capacitance may be $\pm$ 15%, and in such a problem the effect of an anode supply resistance of 10 times the value of R_2 would not be too serious compared with the inherent inaccuracy of the analysis. In the example we have just considered we could make R_2 something like 100,000 ohms and never notice the slightest difference: after all, the components have some losses which are not included in the calculation. Anyway, in a second-order network which we meet very often, L is the leakage inductance of a transformer and we get a path for bias through the secondary, although we can neglect the actual shunt inductances for purposes of calculation.

There is, I think, sometimes a lack of understanding of the reasons why network theorists consider rather artificial networks. It is not that they want to analyse problems which are outside the experience of the practical man, but rather that they are rehearsing situations which illustrate fairly clearly some particular feature. Then when the practical man turns up with his problem, the theorist can say that, stripped of irrelevancies, the system is a network of such-and-such type, and you design it so and so. In these articles we have done a moderate amount of algebra, and we now have this material as background. Let us take a rather different sort of problem, and use our old algebra again.

We shall take a second order filter, and for simplicity we shall take $R_2 = \infty$. That makes $k = 1$, and the response will be given by

$$|N|^2 = 1 + \omega^2(C^2R_1^2 - 2LC) + \omega^4L^2C^2$$

We know that if $C^2R_1^2 = 2LC$ we have a Butterworth response. But suppose we make a practical filter, with some small errors in the component values: what will happen to the response?

Let us take $C' = C + \delta C$ and $L' = L + \delta L$

In this pair of equations, δC and δL are perhaps about 1/10th of the values of C and L respectively.

First, let us assume that the method we use for adjusting the filter is one which makes $\omega^2_0 L'C' = \omega_0^2 LC$ where ω_0 is the design cut-off frequency. We shall have $LC + C\delta L + L\delta C = LC$ (neglecting $\delta L.\delta C$ which is very small) so that $\delta L/L = -\delta C/C$. The two tolerances are equal but in opposite directions. The response is now expressed by

$$|N|^2 = 1 + \omega^2[(C+\delta C)^2R_1^2 - 2LC] + \omega^4L^2C^2$$
$$= 1 + \omega^2[2C\delta CR_1^2 + C^2R_1^2 - 2LC] + \omega^4L^2C^2$$

(dropping terms in δC^2)

The response was to be a Butterworth one, so $CR_1^2 - 2LC = 0$ and we have

$$|N|^2 = 1 + \omega^2\, 2C\delta CR_1^2 + \omega^4L^2C^2$$

Now we know that $C^2R_1^2 = 2LC$ so that we can write

$$|N|^2 = 1 + 2\sqrt{2}.\omega^2\sqrt{LC}.\delta CR_1 + \omega^4L^2C^2$$

If the cut-off frequency is ω_0, with $\omega_0^2LC = 1$, we have

$$|N|^2 = 1 + 2\sqrt{2}\,\frac{\omega^2}{\omega_0}.\delta CR_1 + \omega^4L^2C^2$$

From this, since $CR_1 = \sqrt{2}/\omega_0$ (see Part 1), we reach

$$|N|^2 = 1 + 4\delta C/C.\omega^2/\omega_0^2 + \omega^4/\omega_0^4$$

We could now calculate the shape of the insertion loss characteristics for any particular value of $\delta C/C$. It is quite useful, however, just to look at this expression and to examine what happens if we have $\omega = \omega_0$. Then $|N|^2 = 1 + 4\delta C/C + 1 = 2 + 4\delta C/C$.

Suppose we take $\delta C/C = \pm 0.25$. For $\delta C/C = +0.25$ we have $|N|^2 = 3$ and the response is 4.77db down where it should be 3db down. If $\delta C/C = -0.25$ the response is not down at all at the ideal 3db point, and looking back we see that it has the form

$$|N|^2 = 1 - \omega^2/\omega_0^2 + \omega^4/\omega_0^4$$

This is just the expression we found for the ± 0.625db Tchebycheff case. As we know that Tchebycheff responses are in general more profitable than Butterworth ones, we can aim our design so that it falls between the two by making C′ rather below than above the design value, and L′ on the high side. The simple calculation above has given us a guide as to the way in which we must make provision for the errors of the practical solution.

Suppose, however, that you insist on a Butterworth response at all cost. We have $CR_1^2 = 2L$ so that in the practical case $C'R_1^2 = 2L'$

and therefore $\delta CR_1^2 = 2\delta L$

giving $\delta C/C = \delta L/L$

This time both reactances are in error in the same direction, and again the percentage error is equal. Now the insertion loss characteristic is given by

$$|N|^2 = 1 + \omega^4(L'C')^2$$
$$= 1 + \omega^4(L + \delta L)^2\,(C + \delta C)^2$$
$$= 1 + \omega^4(L^2C^2 + 2L^2C\delta C + 2LC^2\delta L +$$

terms in the products of the errors, which we can neglect).

From this $|N|^2 = 1 + \omega^4L^2C^2\left(1 + 2\frac{\delta C}{C} + 2\frac{\delta L}{L}\right)$

$= 1 + \omega^4L^2C^2(1 + 4\delta C/C)$.

It is not very hard to get from this to the result that if $\omega_0{}^2LC = 1$, the response will be 3db down at $\omega_0(1 - \delta C/C)$. While we quite cheerfully accepted a value of $\delta C/C = -0.25$ before, giving us a response change from ± 1.5db to ± 1.25db, this capacitance tolerance allied with the insistence on a Butterworth characteristic has cost us a 25% change in bandwidth. If we took $\delta C/C = +0.25$ the response would be narrowed by 25%, and would be 6.2db down at the design cut-off, instead of 4.77 db.

We really have collected quite a lot of information from the second-order equation, and now, you notice, it is really practical stuff, which gives us guidance when we are designing a network to be made. We want to lean our design towards the region between a Tchebycheff response and a Butterworth response, and we must adjust the components to give the correct resonant frequencies, rather than just try to get the individual values right.

Let us now look at yet another way in which we can make use of some of our results. The simple filter networks are often used for connecting one valve to another: the circuit used may not look exactly like one of those shown in Table I (page 445, September issue), but it reduces to the filter form if you twist it round suitably. When we have two filter circuits separated by a valve, the overall response characteristic is given by the equation $|N|^2 = |N_1|^2.|N_2|^2$ where N_1 and N_2 relate to the two filter networks. To save labour I shall confine myself to first- and second-order filters, for which

$|N_1|^2 = 1 + x^2$ (first order)

$|N_2|^2 = 1 + ax^2 + x^4$ (second order)

If then we have two first-order filters connected through a valve:

$$|N|^2 = (1 + x^2)(1 + x^2)$$

This is a very dull solution, 6 db down at $x = 1$. If we have a first-order filter and a second-order filter in tandem the most general form we can find for this is:

$$|N|^2 = (1 + x^2)(1 + ax^2 + bx^4)$$
$$= 1 + (1 + a)x^2 + (a + b)x^4 + bx^6$$

I'm not sure if this can be made into a Tchebycheff response: I rather think it cannot. But if we put $b = 1$ and $a = -1$ it reduces to $|N|^2 = 1 + x^6$, a Butterworth response. We have, in fact, obtained a Butterworth response of the third order by combining a first-order characteristic with a second-order Tchebycheff response.

Two second-order filters in tandem will give us, in the most general case:

$$|N|^2 = (1 + ax^2 + x^4)(1 + bx^2 + cx^4)$$

There is an extensive field for study here, but let us put $c = 1$, so that

$$|N|^2 = 1 + (a + b)x^2 + (ab + 2)x^4 + (a+b)x^6+x^8$$

Here we can take $a = -b = \sqrt{2}$, and we are left with

$$|N|^2 = 1 + x^8$$

This particular result is one which happens to crop up in the design of feedback pairs, and we might also find it in the design of a system consisting only of an input circuit, a valve and an output circuit. One valve or four, low-pass or band-pass, its all part of the same family, and one lot of plodding suffices. In this case we should have to go back from

$$|N|^2 = 1 \pm \sqrt{2}\,x^2 + x^4$$

to

$$|N|^2 = 1 + [(CR_p + L\,R_s)^2 - 2LCk]\omega^2 + \omega^4L^2C^2k^2 \text{ (September issue)}$$

and the corresponding equation with k'. We would, perhaps, take k and $k' = 1$, which you can find to be what happens if a pentode is used with an unloaded grid. Then $|N|^2 = 1 + (CR^2 - 2LC)\omega^2 + \omega^4L^2C^2$ and therefore $CR^2 - 2LC = \pm\sqrt{2}\,LC$

$$\frac{CR^2}{L} = 2 \pm \sqrt{2}$$

In this composite equation R is the source impedance for the input filter, and the load impedance for the output filter. C is the valve capacitance, input or output. It doesn't really matter to us which filter we use at the input and which at the output, even though one is a characteristic with a hump in it, while the other is a rounded one. We should check both forms against the valve characteristic in practice, because one presents a much higher load resistance to the valve than the other.

We could, of course, go on to consider what happens if third-order filters are included, but the algebra becomes involved, and is, indeed, sufficiently involved to justify a slightly different approach, based on the distribution of the characteristic frequencies of the networks. I do not propose to discuss this matter at all, because the algebra really does get rather beyond the permitted limit.

Nothing in this article, of course, will enable you to design a filter wihtout thought. None of the equations should be taken on trust, because they are only introduced as examples of how to set about the job, and I usually work them out from first principles every time I need them. But if you can work out what happens in a simple RC circuit, you can calculate a second-order filter, too. If you can do that, you can try the third-order filter: better still, you can try the problems of variations, like the effects of component tolerances we have been considering above, or problems connected with the phase characteristic, which I haven't discussed at all. Then there's quite a different set of response characteristics, of the general form $N = (1 + j\omega/\omega_0)^n$. These are thoroughly well damped, and give no overshoot at all when a step wave is applied to them. Some of this work, perhaps, is just shadow boxing, just a light paddle between the locks, a chess problem by the fireside; but it is all part of the process of building up a solid foundation on which further development can take place.

NEW EPOXIDE RESIN

IMPROVED strength and toughness is claimed for a new epoxide resin ("Epophen" M-777) made by Leicester Lovell and Company, North Baddesley, Southampton. It is used as a contact adhesive for bonding metals, glass and plastics, as well as wood and other materials which are normally glued. In addition it can be used as a casting resin, and one application is in the production of press tools for sheet metal working Castings up to ½-ton in weight are possible, one reason being that there is no appreciable evolution of heat during the setting process.

For test purposes, a kit containing resin, catalyst and filter, with means of measuring and mixing and full instructions, is obtainable from the manufacturers, price 7s. 6d.

High-Quality Tape Recording

Circuit Design for a Recorder Using High-impedance Heads

By A. F. FISCHMANN*, A.M.Brit.I.R.E.

THE design described in this article is simple and may be adopted easily to different tapes and high impedance recording heads. Frequency-selective feedback is used throughout the system in order to modify the frequency response.

The author's equipment consists of the Model 5 Tape Desk built by Bradmatic Ltd., Station Road, Aston, Birmingham 6. It is equipped with erasing, recording and reproducing heads, all of the high-impedance type. Its performance using MC1-111 tape of the Minnesota Mining and Manufacturing Company Ltd. at a speed of 7.5in/sec. was found to be as follows:

Response within $\pm$3db between 40-10,000 c/s.
Signal-to-noise ratio 45 db.

The distortion was not measured but the reproduction of high-quality discs through the recorder proved to be indistinguishable from the original. An extension of the frequency range could have been readily obtained, using additional equalization.

The system consists of a recording amplifier, playback amplifier and supersonic bias oscillator and may be used in connection with any high-quality pre-amplifier as a front-end control unit for a power amplifier. Obviously all the equalization necessary for the truthful reproduction of records or any other programme must be provided within the pre-amplifier, and the system itself is designed for flat response within the above-stated limits between its input and output. The main amplifier should provide the conventional tone control circuits (treble and bass lift and cut). A block diagram of the complete system is shown in Fig. 1 together with the selector switch. Switch S_1 is provided in order to compare the performance of the unit with direct transmission and is especially useful for the adjustment of its frequency response in the absence of accurate measuring gear.

The diagram of the record amplifier is shown in Fig. 2. To suit the special requirements of a tape recorder, it should provide a constant current through the recording head at all transmitted frequencies, thus ensuring equal magnetization of the tape over the whole transmission band. In addition, treble lift may be necessary to compensate for losses due to the tape and the airgap of the recording head, whose width sets a definite limit to the highest transmitted frequency. The constant-current characteristic is frequently achieved by providing a frequency response rising by 6db per octave. The signal is then applied to the recording head which represents a mainly inductive load and a substantially constant-current characteristic is thus obtained. Different arrangements use a resistance of about 50kΩ through which the recording head is connected to a conventional amplifier. The latter has to provide a considerable output voltage at all frequencies due to the losses in that resistance and may give rise to distortion if not properly designed.

In this amplifier a different way was chosen. The output stage, Fig. 3(a), will be considered separately and its equivalent circuit diagram is shown in Fig. 3(b), where

$$R'_i = R_i + R_k(\mu + 1) \qquad (1)$$

(R_i = internal resistance of V2).

The voltage across the load

$$E_L = -\frac{\mu E_{in} Z_a}{R_i' + Z_a} \qquad (2)$$

$$\text{where } Z_a = \frac{R_a Z_L}{R_a + Z_L}$$

$$g'_{m2} = \frac{\mu}{R_i' + Z_a} \qquad (3)$$

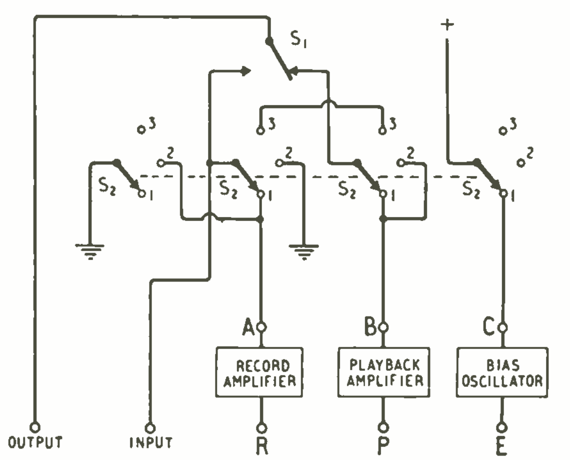


Fig. 1. Block diagram of the system Switch positions are: 1, recording: 2, reproducing: 3, direct.

* Israeli Ministry of Defence, Scientific Department.

Symbols

R_{in}	=	Internal resistance of valve *n*.
R_k	=	Cathode resistance.
R_{an}	=	Anode resistance of valve *n*.
μ	=	Amplification factor.
g_{mn}	=	Mutual conductance of valve *n*.
g_{pn}	=	$\frac{1}{R_{in}}$
G_k	=	$\frac{1}{R_k}$
G_a	=	$\frac{1}{R_a}$
Z_{out}	=	Output impedance of recording amplifier.
i_r	=	Recording current, C_{02} connected.
i_r'	=	Recording current, C_{02} disconnected.
L_R	=	Inductance of recording head.
r_R	=	Equivalent series resistance of recording head.

System

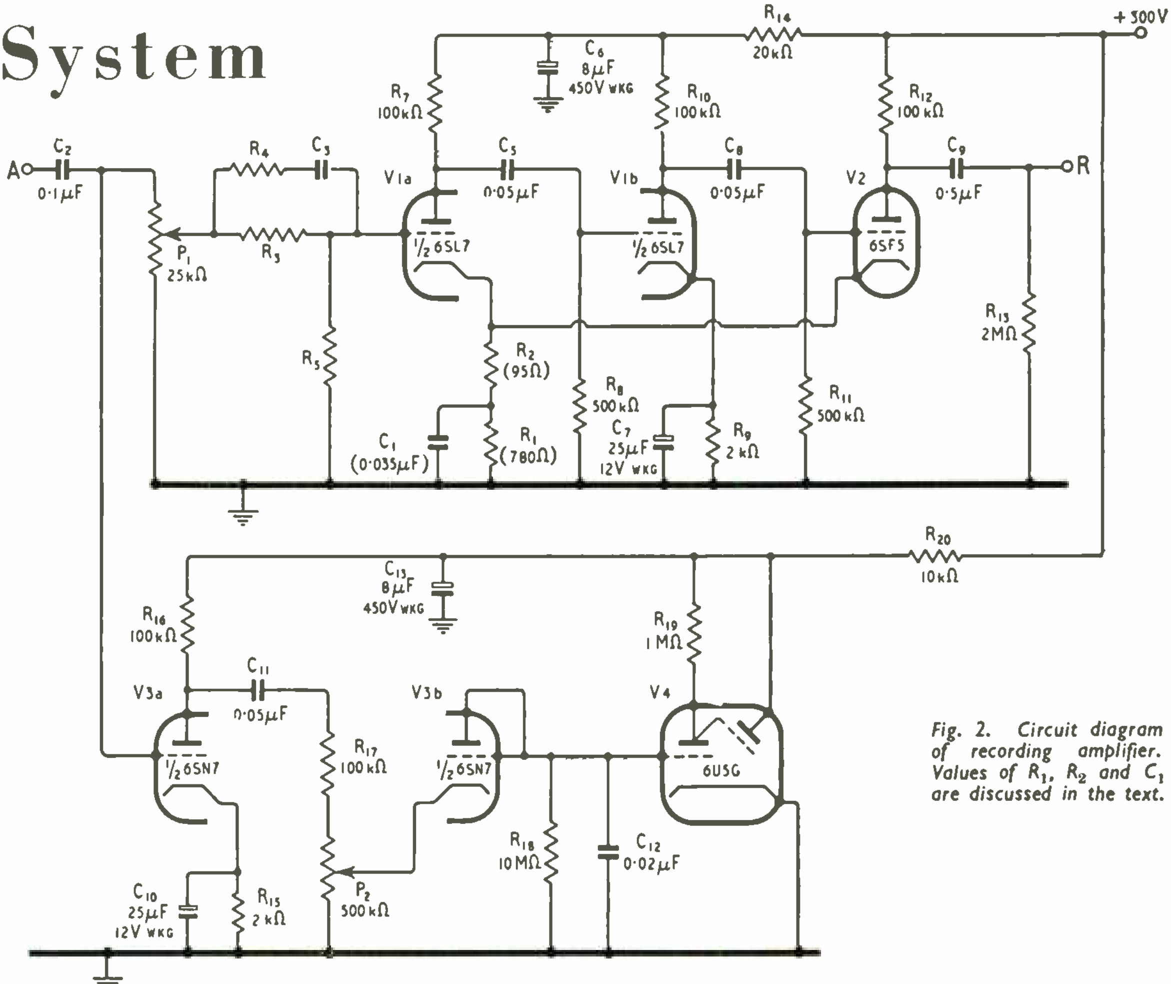


Fig. 2. Circuit diagram of recording amplifier. Values of R_1, R_2 and C_1 are discussed in the text.

may be considered as the effective transconductance of the valve V2 due to the current feedback from the cathode. As $R_i' \gg Z_a$ within the transmitted frequency spectrum, $g'_{m2} \approx \frac{\mu}{R_i'}$ and V2 is working as a constant current generator.

R_k may be designed as to make R_i' sufficiently high in order to drive a recording head with a constant current over a reasonable frequency band. However, in that case V2 would need a driving voltage of approximately 20 volts, which can hardly be provided by any of the pre-amplifiers generally in use. Therefore a double triode is added and the feedback is extended over two additional stages. Using the simplified diagram shown in Fig. 4 and putting $G_{a1}=G_{a2}=G_{a3}=0$, the output admittance of that circuit is

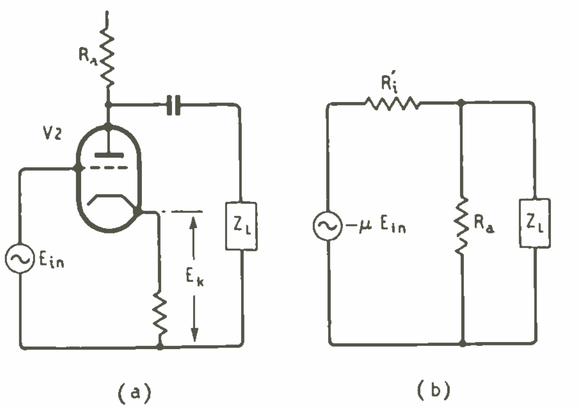


Fig. 3. (a) Output stage of recording amplifier and (b) its equivalent circuit.

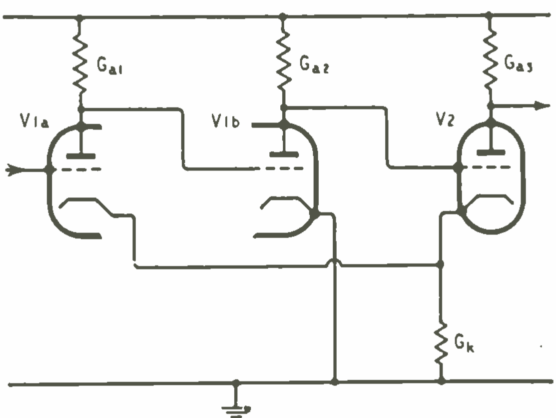


Fig. 4. Simplified diagram of feedback in the recording amplifier.

$$Y_{out}=g_{p2}\left(1-\frac{1}{1+\beta G_k}\right) \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

$$\text{where } \beta=\frac{1}{g_{m2}+g_{p2}+\frac{(g_{m1}+g_{p1})g_{m2}g_{p2}}{(g_{p1})^2}}$$

and g_{m1}, g_{m2}, g_{p1}, g_{p2} are the respective mutual conductances and plate conductances of V1 and V2.

As $\beta G_k \ll 1$, therefore $1-\frac{1}{1+\beta G_k}\approx \beta G_k$ and $\frac{1}{Y_{out}}=Z_{out}\approx\frac{1}{g_{p2}\beta G_k}=\frac{14.4\times10^6}{g_{p2}G_k}$ MΩ for practical values of g_m and g_p.

Z_{out} may therefore be neglected in comparison with R_a and the latter determines solely the output impedance of the complete amplifier. Its current amplification will be inversely proportional to R_k and may be made to rise over a certain frequency range by introducing an R−C network (R_3, R_4, R_5, C_3, in Fig. 2). It should be designed to compensate for the loss of high frequencies as already mentioned, and may extend the flat frequency range by one octave. It should be noted that the high-frequency response of the amplifier may be controlled by extremely low impedances, avoiding the conventional high-impedance type networks which are most liable to capacitive pick-up if not properly screened. The recording amplifier response may also be easily adapted to different kinds of tape speed by changing C_1, which will move both the low and the high frequency turn-over point by an equal amount. The value of these components should be determined experimentally according to the characteristic of the individual tape and recording head. Their design procedure will be outlined later in this article.

A magic eye V4 is connected to the input of the recording amplifier through V3. It should close for full modulation of the tape and its sensitivity may be adjusted by P_2 (Fig. 2).

The playback amplifier (Fig. 5) is designed along conventional lines. A signal recorded with a constant current characteristic will provide an output rising by 6db per octave at the terminal of the playback head. In this particular case the response was measured as shown in Fig. 6, using MC1-111 tape at a speed of 7½in/sec. It rises by 6db per octave up to a frequency of 1,600c/s, then flattens out gradually and drops considerably above 6,000c/s. This is caused by losses due to self-demagnetization, penetration and the gap effect[1].

The response of the playback amplifier should be inverse to the curve shown in Fig. 6 and should therefore fall by 6db per octave from 40c/s per second (the lowest transmitted frequency) up to 2,500c/s. This is controlled by the frequency-selective feedback network between the anodes of V5 and V6 consisting of $R_{53}+R_{54}$, R_{56}, C_{52}, R_{55}. The value of $R_{53}+R_{54}$ will determine the high-frequency turnover point, but leave the low-frequency turnover point unchanged as long as the other values of the network are retained. The playback amplifier characteristic may therefore be easily adapted to different tape speeds by changing the value of $R_{53}+R_{54}$.

C_{51} belongs to the de-emphasis network and will be mentioned later. The respective values of these components are not specified and the reader is referred to the alignment procedure as outlined at the end of this article.

As the output voltage at the playback head is extremely low, especially at low frequencies, the amplifier should be designed for minimum noise level. A triode-connected 12SJ7, with the heater d.c. fed, preferably from the main amplifier anode current, is used in the first stage, with an equivalent noise resistance due to shot effect of approximately $\frac{2.5}{g_m}$ ohms at the input[3]. To this is added the flicker effect which increases with the square of the emission current, but decreases with the inverse square of the frequency[4]. Because of this frequency dependence, the flicker effect becomes appreciable at low frequencies where the amplification of the playback amplifier is just at its maximum value. The noise due to the flicker effect may however be held sufficiently low reducing the anode current of V5 by making R_{57} 0.3MΩ.

Finally we come to the erase and bias oscillator (Fig. 7). Greatest possible freedom of harmonics is desired to ensure low noise level and low distortion of the recorded signal. It consists of a cathode follower type oscillator, the principle of which was described elsewhere[2]. The variable feedback resistor P_3 provides a method of controlling the negative resistance injected into the tuned circuit with great accuracy. It may be adjusted to provide an amplitude of 8 volts at the cathode of V7. Under those conditions the grid-cathode voltage of that valve is small enough to ensure its linear operation, and such an oscillator will therefore provide an output containing less har-

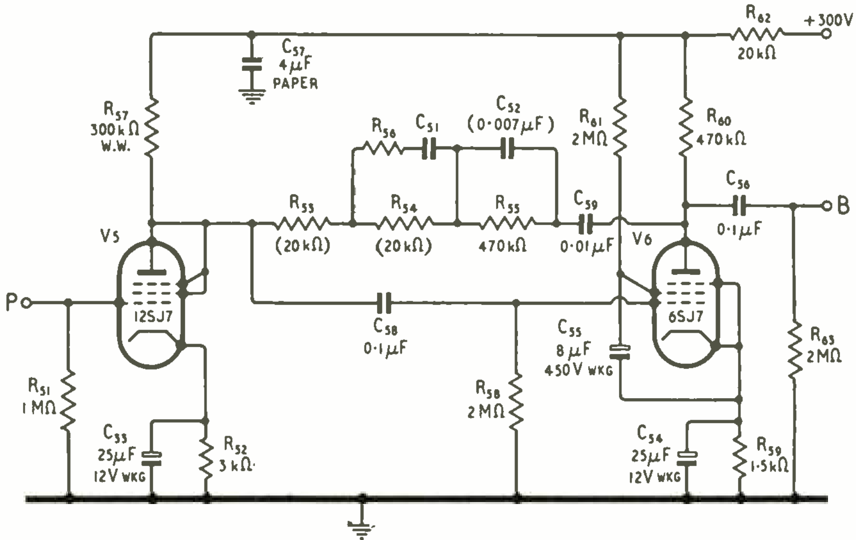


Fig. 5. Circuit diagram of playback amplifier. Values of R_{53}, R_{54} and C_{52} are discussed in the text. The heater of the first valve should be fed with d.c., preferably from main amplifier anode current.

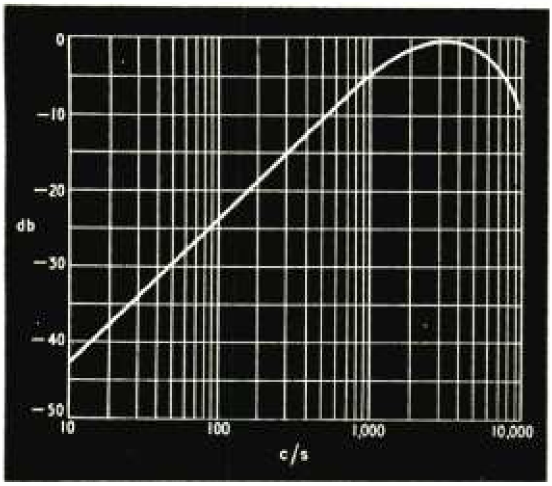


Fig. 6. Measured response at playback head, with constant-current recording.

monic distortion than a conventional one. However, its amplitude is not absolutely constant over a considerable time and therefore P_4 and M (Fig. 8) provide a means of adjusting the bias amplitude accurately during recording. The erasing head in the author's equipment was found to have an inductance of 29mH and an equivalent series resistance of 4,700 ohms at 60kc/s. It is brought to series resonance with C_{90} (Fig. 8) thus providing the necessary load of 5,000 ohms for V8. In addition, the output voltage is stepped up, the Q of the circuit being about 2. This selectivity will also provide some attenuation of harmonics generated in the output valve.

It is important that C_{90} should be connected directly to the erasing head thus leaving the capacitance of the connecting cable in parallel with the tuned anode load of V8, which should resonate at the oscillator frequency of 60 kc/s.

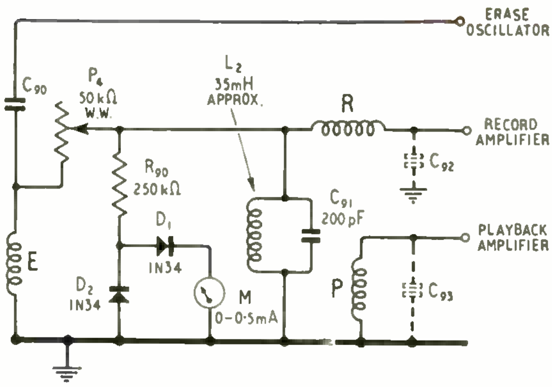


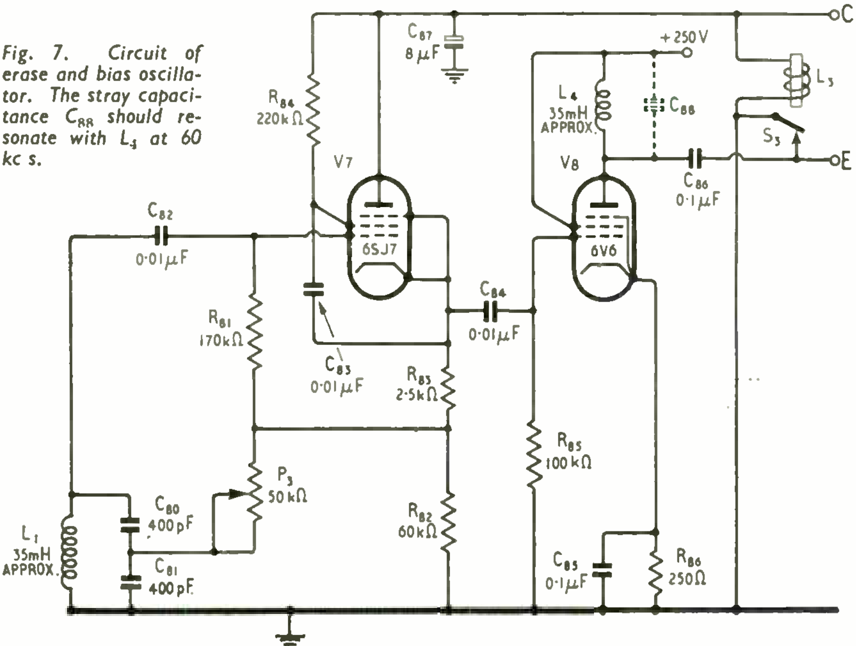
Fig. 8. Connections of magnetic heads and bias level meter. R, recording head: P, playback head: E, erase head. C_{92} and C_{93} are the capacitances of the connecting leads. C_{90} is adjusted to resonate with E at 60 kc/s.

The anode of the oscillator valve V7 is bypassed to earth with 8μF and consequently its anode voltage will decrease only gradually after interruption of the supply voltage by S_2, which ensures slow damping out of the oscillations. This is important in order not to leave any magnetism in the erase and recording head after the oscillator is switched off. The relay L_3, S_3 serves to short the erasing head to earth when the oscillator is inoperative. It should close only after the oscillations are completely damped out. This is accomplished by removing the copper rivet generally used in conventional relays. This rivet serves to prevent sticking of the relay due to remanent magnetism, after the current is interrupted. With this small alteration the relay will open with a certain delay and thus provide the necessary characteristic.

The recording bias is taken from the erasing head and is adjustable by P_4 (Fig. 8). The parallel tuned circuit L_2C_{91} resonates at 60 kc/s, thus forcing the bias current mainly through R and C_{92} (the capacitance of the connecting cable), the voltage drop across the latter being negligible. The filter formed by P_4 L_2 C_{91} will also provide additional attenuation of harmonics of the bias frequency.

The situation is different at recorded frequencies: L_2 provides then a short circuit and C_{92} will resonate with the recording head R somewhere above the highest transmitted frequency. This method of coupling was chosen in order to prevent the bias from entering the record amplifier and at the same time to maintain the high input impedance of R as seen by the record amplifier. Any capacitive load of the latter would considerably impair the response at the highest audio frequencies due to its high output impedance.

Fig. 7. Circuit of erase and bias oscillator. The stray capacitance C_{88} should resonate with L_4 at 60 kc s.



A value of 200pF was chosen for C_{92} to resonate with the recording head used in the writer's equipment at 14,000 c/s. The voltage drop across that capacitance due to the bias will be fed into the record amplifier and, although small, it may give rise to beats with the recorded frequencies. It will, due to the feedback action of the amplifier, appear in antiphase at the grid of V2 with an amplitude determined by the voltage divider consisting of the cathode resistor and the internal resistance of V2. In order to keep this amplitude low a valve with a high internal impedance should be chosen for V2 and a 6SF5 having an internal resistance of about 60 kΩ was found to be adequate. However a pentode with an internal resistance of the order of a megohm may be chosen in extreme cases.

Circuit Adjustments

The values of the various components determining the frequency response should be found by trial and error. A signal generator providing frequencies between 40-15,000 c/s and a valve voltmeter calibrated in db are highly desirable. First the amplifier will be operated with C_1, C_3, C_{61} and C_{62} disconnected. The optimum bias may now be adjusted by means of P_4 using a signal of about 3,000 c/s. When increasing the h.f. bias from zero, the volume will rise and distortion will decrease considerably, until a maximum of volume is reached. A further increase of the bias will cause additional reduction of distortion but also of the volume. The optimum setting of P_4 for minimum distortion will therefore slightly increase the signal-to-noise ratio, but minimizing the distortion is considered to be of primary importance. In the writer's equipment a bias of 135 volts across the recording head was found to correspond to minimum distortion.

After adjusting the bias, the amplification of the record amplifier may be adjusted by means of the feedback resistor $R_1 + R_2$ (see Fig. 2) to deliver at the recording head a signal as specified by the makers of the tape desk. The response of the playback amplifier should now be plotted, yielding a curve similar to the one shown in Fig. 6. Then, taking the flat response between 2,500-4,000 c/s as the reference level, the value of C_{62}, $R_{63} + R_{64}$ may be found for flat response down to 40 c/s. $R_{66}C_{62}$ determines the low-frequency turnover and $R_{63} + R_{64}$, C_{62} the high frequency turnover. Now the treble boost provided by C_1, R_1, R_2 in the record amplifier may be adjusted to extend the frequency response at the high-frequency end. The time constant R_1C_1 determines the low-frequency turnover point and R_2C_1 the high-frequency turnover point.

In addition the response of the record amplifier at the high frequencies depends on the value of C_{92}, which works out as follows:

$$i_r \propto \frac{1}{R_{a2}(1-\omega^2 L_R.C_{92})+r_R+j\omega L_R\left(1+R_{a2}\dfrac{C_{92}.r_R}{L_R}\right)} \quad (5)$$

and the ratio of the response with C_{92} connected to the response without C_{92}

$$\frac{i_r}{i_r'} = \frac{R_{a2}+r_R+j\omega L_R}{R_{a2}(1-\omega^2 L_R.C_{92})+r_R+j\omega L_R\left(1+R_{a2}\dfrac{C_{92}.r_R}{L_R}\right)} \quad (5a)$$

Consequently the response of the playback amplifier may be lifted at the high frequencies by a ratio indicated in equation 5a.

Additional treble equalization may be obtained by tuning the playback head to resonance at the high end of the transmitted frequency band. The capacities C_{92}, C_{93} may be fully or partly realized by the capacity of the connecting cables. An effective means of improving the signal-to-noise ratio is to use pre-emphasis for frequencies above 1000 c/s similar to the characteristic used for records. This may be introduced by the conventional network R_3, R_4, R_5, C_3 connected to the input of the recording amplifier†. The loss of amplification due to this network may be compensated by decreasing the feedback from R_1, R_2.

As R_1, R_2, C_1 have been determined previously, the time constant R_1C_1 and R_2C_1 should remain unchanged in order not to affect this adjustment. The de-emphasis is incorporated in the feedback network of the playback amplifier and consists of R_{63}, R_{64}, R_{66}, C_{51}.

If records are played with a magnetic pickup (giving constant output from constant velocity recording), they may be recorded on the tape without any correction of the frequency band above 1000 c/s, which will provide directly the wanted pre-emphasis characteristic.

The only correction applied in that case will then be the de-emphasis, incorporated in the playback amplifier.

† See, for example, "Radio Designers Handbook," by E. Langford Smith, p. 653.

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3 "Vacuum Tube Circuits," by L. B. Arguimbau, Chap. 3, 15. (Chapman and Hall).

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F.M. Car Radio covering 87-109 Mc/s with a.f.c. for locking on to signal and 2μV sensitivity. The aerial is embedded in a strip of plastic for sticking to inside of windscreen. Descriptive leaflet from Hastings Products, 171 Newbury Street, Boston 16, Mass., U.S.A.

Prefabricated Cabinets and telescopic mountings; a new illustrated catalogue including additions and improvements to the Widney-Dorlec range of parts. From Hallam, Sleigh and Cheston, Bagot Street, Birmingham 4.

Tape Table made by Lane with two speeds and frequency response of 40 c/s-10 kc/s. Tape reels are locked when machine is off to avoid tape spilling. Leaflet from Verdik Sales, 8 Rupert Court, Wardour Street, London, W.1.

Flexible Shafts; a detailed and well-illustrated handbook describing their construction and characteristics, with graphical and tabular data, and explaining their application to various remote control problems. From The S. S. White Co. of Great Britain, 126 Great Portland Street, London, W.1.

Audio-frequency Amplifiers; leaflet describing the "Astronic" 12-watt range, for various inputs, from N. Miers and Co., 115 Gower St., London, W.C.1.

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More Rectification

By "CATHODE RAY"

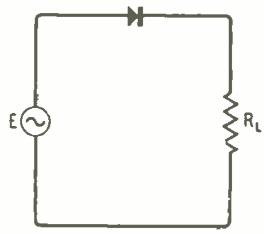
Far-reaching Effects of a Reservoir Capacitor

LAST month we considered Fig. 1, the simplest possible rectifier circuit, reduced to its simplest possible terms, namely a resistanceless generator of pure sine waves feeding a pure resistance load through a perfect rectifier, i.e., one having no resistance at all to current in one direction and infinite resistance to current in the opposite direction. We considered the readings of perfect voltmeters of various types connected to read the three possible voltages in the circuit—across generator, load, and rectifier—and perfect ammeters to read the one possible current. And in spite of taking a long time over this apparently simple and straightforward job, we didn't even finish it. We drew up a table of readings given by three types of voltmeter: (1) the ordinary moving-coil type used for d.c., which responds to mean values; (2) the same with a full-wave rectifier to adapt it for a.c., which responds to mean values of rectified voltages, but is scaled to read r.m.s. values, which are 11% higher; and (3) the electrostatic or moving iron or any other type that actually responds to r.m.s. values as well as being scaled in them. Most valve voltmeters are outside all these categories because, although usually scaled to read r.m.s. values, they respond to peak values. Our first job now is to complete the table by filling in the entries for this type of meter.

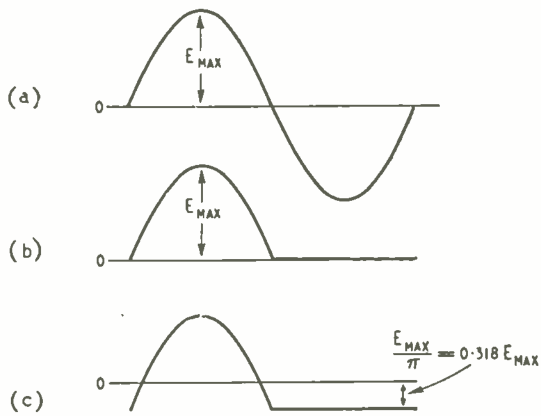
Again let us assume that the instrument is perfect, so that the rather complicated matter of its errors doesn't arise. Now we know that the r.m.s. value of a sine wave is equal to its peak value divided by $\sqrt{2}$; in the usual symbols, $E = E_{max}/\sqrt{2} = 0.707E_{max}$. So to make the peak voltmeter read r.m.s. values directly, it is arranged so that it indicates 0.707 times the peak value. Consequently when connected across the generator in Fig. 1, where it sees waveform 2(a), it reads E, the r.m.s. value—like the other types of a.c. voltmeter.

Next, connect it across the rectifier or the load. In either of these positions it sees waveform 2(b). If the voltmeter has no blocking capacitor or transformer coupling it responds to the peak voltage from the zero line. Connected one way round, this peak voltage is E_{max}, and as the instrument is scaled to read 0.707 times this it reads E as before. If connected the other way round it receives only the half-cycle that has been removed by rectification, so the reading is nil.

But most valve voltmeters have a series capacitor,



Right: Fig. 1. The simplest possible rectifier circuit again.



Below: Fig. 2. Waveforms associated with Fig. 1: (a) across the generator; (b) across rectifier or load resistance; (c) after using a blocking capacitor on (b) to exculde the d.c. part.

TABLE I

Type of voltmeter		Across generator (with or without blocking capacitor)	Across R_L or rectifier (without blocking capacitor)	Across R_L or rectifier (with blocking capacitor)
Mean (moving-coil)		0	0.450E *or* —0.450E	0
Rectifier ..	calibrated to read r.m.s. values of sine wave	E	0.500E	0.551E
Square-law ..	calibrated to read r.m.s. values of sine wave	E	0.707E	0.545E
Peak.. ..	calibrated to read r.m.s. values of sine wave	E	E *or* 0	0.682E *or* 0.318E

which blocks the d.c. component, and after the surge caused by the charging of this capacitor at the moment of connection has died away the voltmeter sees waveform 2(c). We have already found that the d.c. component is equal to the peak value divided by π, and the voltages are in the same proportion. So if connected one way the peak value is $(1-0.318)$ E_{max}, and the other way $0.318E_{max}$; and the readings, being 0.707 times these, are 0.682E and 0.318E respectively.

So except across the generator, where the waveform is symmetrical, the reading of the peak voltmeter depends on which way it is connected; hence the two pairs of entries in Table I for this type. The only other reading that is affected by direction is the mean value, and that is not with regard to the actual value of the reading but only its polarity.

We must remember once more that if the generator provided some other waveform than sinusoidal, most of the figures in the table would be different. We noted before that if a rectifier meter were connected across a source of square waves (whose mean, r.m.s. and peak values are all equal) it would read 11% high —because it is designed to do this in order to allow for the inequality of the mean and r.m.s. value of sine waves. The inequality between *peak* and r.m.s. values is the other way around and greater; our compensated peak voltmeter would read square waves nearly 30% low.

Now for Current

No type of ammeter that I know reads in proportion to peak values. The only way would be to put a very low resistance in circuit, amplify the voltage across it, and measure that with a peak voltmeter. A cathode-ray oscillograph would do as the amplifier and voltmeter. Any such arrangement would hardly be calibrated in either r.m.s. or mean values, so would not be comparable with other types of current meter. But while we are at it we might as well tabulate last month's findings for these other types (Table II).

Fig. 1 being a purely series circuit, there is only one current throughout, the waveform of which is as Fig. 2(b). But in the current table there are two columns of readings, because some meters receive the current directly, while others are coupled by a transformer, which eliminates the d.c. component. The moving-coil d.c. meter is always directly connected, for obvious reasons. But the rectifier meter can be connected either way, as in Fig. 3, and when it is used in a rectifier circuit the reading depends on which. In practice it is almost always transformer-coupled (b), because this allows the range to be varied by varying the number of primary turns in circuit. Fig. 3(a) only provides one range, up to the maximum rating of rectifiers and meter, because owing to the varying resistance of the rectifiers the range cannot be varied in the usual way by shunts. Lastly the square-law type, which is usually a heat operated instrument, because that can be accurately calibrated on d.c. The point to remember is that the true r.m.s. reading which it indicates (regardless of waveform) is, with a sinusoidal input to this simple circuit, 1.57 times the mean current, as read on a d.c. meter, so its heating power is $1.57^2 = 2.46$ times as much.

Having claimed so much of your valuable time on this absurdly simple(?) circuit, I may be running a grave risk of assault if I now calmly announce that it is of theoretical interest only, being rarely used in practice. But I hope that the effort may be seen to have been worth while, as a convincing warning against tackling even the simplest rectifier circuits and calculations without due care and attention. They are thoroughly deceptive things. That being understood, however, perhaps I can deal with the practical types more sketchily.

The reason why Fig. 1 is rarely used in that simple form is that d.c. in the Fig. 2(b) form is rarely desired. But the smoothing circuits needed to reduce it to pure d.c. have the incidental effect of radically altering the rectifier circuit. So let us now look into rectifier circuits modified by smoothing.

The simplest is Fig. 1 with a capacitor connected across R_L, as in Fig. 4(a). A closely related variety is Fig. 4(b). These come in for a great deal of attention, because they are so much used; nearly all detectors, nearly all valve voltmeters, and nearly all the rectifier circuits in a.c./d.c. sets are essentially one or other of these types. Quite recently they were discussed at some length from the valve voltmeter point of view.* Pages and pages of data appear on them in the *Radio Designer's Handbook* and indeed in most books on radio. So all I am going to do now is to outline how C in Fig. 4(a) affects the currents and voltages.

If we were to continue on our hitherto ideal lines, we would assume that C was infinitely large, so as to provide perfect smoothing. But these assumptions would lead to the generator being required to supply an infinitely great current for an infinitesimally small time during each cycle; and to avoid such arrant

* M. G. Scroggie, *Wireless World*, June, p. 284, and July, p. 339, 1954.

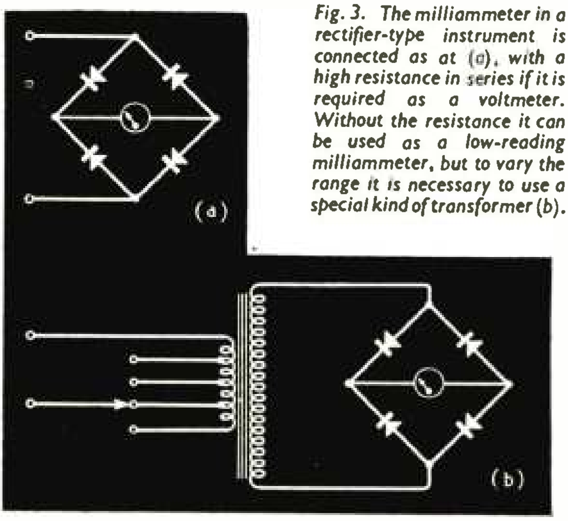


Fig. 3. The milliammeter in a rectifier-type instrument is connected as at (a), with a high resistance in series if it is required as a voltmeter. Without the resistance it can be used as a low-reading milliammeter, but to vary the range it is necessary to use a special kind of transformer (b).

TABLE II

Type of ammeter		Directly in circuit	Transformer coupled
Mean (moving-coil)		I_{av}	0
Rectifier	Calibrated to read r.m.s. values of sine wave	1.11 I_{av}	1.22 I_{av}
Square-law	Calibrated to read r.m.s. values of sine wave	1.57 I_{av}	1.21 I_{av}

Peak value $= \pi I_{av}$

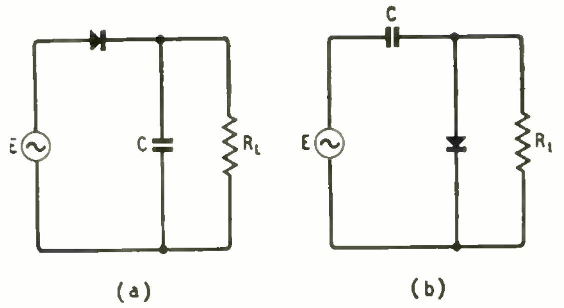


Fig. 4. Two varieties of the reservoir type of half-wave rectifier. (a) is used for power supply, and either (but usually (b)) for valve voltmeters.

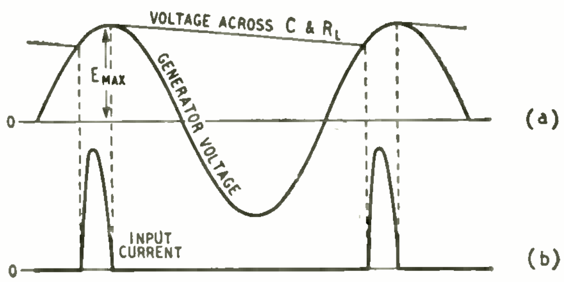


Fig. 5. Voltage and current waveforms for the Fig. 4 type of rectifier circuit.

nonsense we have to acknowledge the existence of resistance in the generator and rectifier (conducting phase). And this leads to all the complicated data and calculations referred to. However, the general picture is that when the rectifier conducts C charges up to nearly E_{max}; and, provided its capacitance is large enough to keep current flowing steadily through R_L during the rest of the cycle, its voltage remains not far short of E_{max} (Fig. 5(a)). Consequently it is only close to the peak that the generator voltage is great enough to overcome the capacitor voltage and make current flow through the rectifier. During this small fraction of the cycle, the generator has to supply enough current to C (Fig. 5(b)) to keep current going through R_L all the time. The larger C is the less its voltage drops below E_{max}, the shorter the time during which current enters it from the rectifier, and the heavier that current must be. It is quite usual for it to be five or even ten or more times the load current. In fact, in power circuits, having to supply a lot of load current, it is necessary to limit C, or to insert extra series resistance, or both, if the rectifier is to be preserved from an early death.

What about the meter readings and the actual currents and voltages? In valve voltmeters and detectors, where it is output voltage rather than current that counts, the resistance R_L can be made very high, and the load current consequently very low; so the output voltage is nearly steady at only a very little below E_{max}. (This is very different from Fig. 1, where the output voltage (Fig. 2(b)) is very unsteady and only momentarily reaches E_{max}, its average value being less than one third as much.) In these circumstances the output current is of course also steady at nearly E_{max}/R_L. But the generator current through the rectifier, although it is bound to have the same average value (as would be shown by a moving-coil milliammeter), consists of a series of brief pulses. If these pulses were square-cut, then if they lasted one *n*th of each cycle they would be *n* times the load current. But being peaky their peak value is even greater than this. Almost the only practical way to measure it is by means of the c.r. oscilloscope across a low series resistance.

As you may have guessed from our experience with the Fig. 2(b) waveform, the true r.m.s. value of this current is much greater than its mean value. Suppose we take as an example a circuit in which the mean output voltage is 95% of E_{max}. Fig. 30.5 in *Radio Designer's Handbook* shows that if the series resistance (generator, rectifier, and any added) is 0.0005 R_L, the value of ωCR_L (time constant of the load multiplied by 2π times the frequency) is about 37. Fig. 30.8 in the same book then shows that (assuming constant rectifier forward resistance) the peak input current is nearly 20 times the mean output current, and the r.m.s. input current is about $3\frac{1}{2}$ times the output. So although a d.c. milliammeter in series with the rectifier would indicate the same current as in R_L, its heating effect in a given resistance would be about $3\frac{1}{2}^2$ or 12 times as great! If one chose the gauge of wire for the transformer secondary (acting as the generator) on the basis of the d.c. meter reading, one would probably be able to burn one's fingers on the transformer after it had been running some time. That is if the rectifier, chosen on the same basis, was still rectifying, which would be unlikely. For power units it is uneconomic to attempt a 95% voltage yield; it is more usual practice for the output voltage, after allowing for the drop in further smoothing if any, to be equal to about E, say 70% of E_{max}. The peak current is then of the order of five times the output, and the r.m.s. value a little over twice. Even so, its heating value is some five times that of the same current after smoothing.

Readings to Take

The measurement procedure with this type of rectifier circuit, then, is to use an ordinary d.c. milliammeter for the output current. Multiplying this by R_L gives the mean output voltage. Comparing this with the peak input voltage (which is $\sqrt{2}$ times its measured r.m.s. value if sinusoidal waveform can be assumed) and knowing the component values, we can look up data sheets to find the r.m.s. and peak input currents, which are needed to design the transformer and choose the rectifier. If the data sheets are lacking, or we want to check them, it is necessary to have a true r.m.s. current meter (preferably a thermojunction type; but keep it shorted except when taking a reading, or it will almost certainly burn out when switching on!) and the low resistance and oscilloscope or sensitive peak voltmeter. A rectifier type of a.c. milliammeter is really more than useless almost anywhere in a rectifier circuit, because the readings it gives are quite different from what they purport to be. A rectifier voltmeter can be used to measure the generator voltage, provided pure waveform can be guaranteed; but even quite small generator impedance is enough to upset this guarantee, because of the very peaky current waveform.

Usually there are some additional smoothing components between C and R_L, and these affect the situation in rather a complicated way. But provided that C is relatively large, the basic action of the circuit as already described is not altered out of all recognition. Apart from the actual smoothing effect, which I discussed in the October and November 1949 issues, the main practical point to reckon with is the voltage drop in the series smoothing components, and that is easy enough.

Although Fig. 4(a) is very widely used for power supplies, that is because it is almost the only choice in a.c./d.c. sets, and not because it is a good circuit for power supplies. Because the rectifier passes current only once per cycle, the current it does pass has to be so large that a high-rated rectifier must be used. And for the same reason the voltage drop between current pulses is apt to be large, and the output needs a lot of smoothing. When the source is definitely a.c., so that a transformer can be used, it nearly always is used, in order to take advantage of certain benefits obtainable therewith. One of the benefits is the ability to use both half-cycles of the source, replenishing the reservoir C twice per cycle. Another is the ability to step up the voltage, not only in the transformer itself but also by the arrangement of rectifiers. There is the centre-tapped transformer full-wave circuit, the rectifier-bridge full-wave circuit, two sorts of voltage doubler, and a voltage quadrupler, besides some rarer varieties. The first of these is the commonest, and the only one I am going to take for the present.

Fig. 6 is the basic circuit. I have drawn it that way, because it is how it usually appears in circuit diagrams; but obviously it would look simpler and clearer if C and R_L were drawn straight across between the transformer centre tap and the junction between the rectifiers. That would help to bring out the fact that it is really the same as Fig. 4(a) supplemented by a second rectifier fed by a source in opposite phase, to double the number of current input pulses per second. But except for the reduction in output voltage fall during the cycle, there is no increase in voltage obtained by the doubling of the end-to-end secondary voltage. For this reason it is not the circuit one would choose for high-voltage low-current output. Even an ordinary receiver requiring, say, 350 V., necessitates a considerably higher transformer voltage. Using our rough rule that in power units the output voltage is about equal to the r.m.s. input voltage, we see that the peak voltage of each half of the secondary is $\sqrt{2} \times 350 = 495$ V., so the total peak secondary voltage is practically 1,000—much too high for one's health if one gets it between the hands!

Assuming that the circuit is balanced—rectifiers and secondary half-windings identical—it is obvious that if we connect our moving-coil milliammeter in series with either rectifier it should read half the load current. In practice there is usually an appreciable unbalance, but it ought not to be more than 10 or 15 %. As with the half-wave circuit, the number of times the r.m.s. and peak values are greater than the mean depends on the component values and frequency. Approximately the same ratios—and output-voltage ripple—as in the half-wave circuit will be obtained in a full-wave circuit having everything the same, except for rectifier, etc., resistance being doubled on each side (because each carries only half the current) and C halved (because the ripple frequency is doubled).

Note that, although the voltage of the whole secondary coil in the full-wave circuit is in the same direction from end to end, so that the a.c. is likewise, the d.c. flows in opposite directions in the two halves.

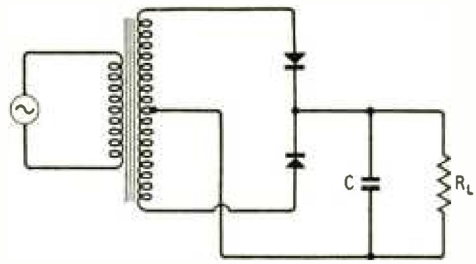


Fig. 6. Modification of Fig. 4(a) for full-wave rectification. This is the commonest type for supplying h.t. in a.c. receivers.

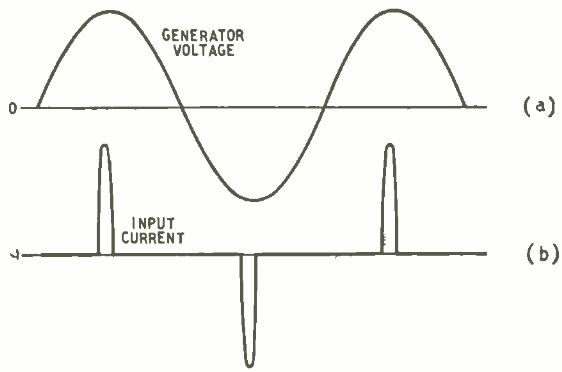


Fig. 7. Voltage and current waveforms in Fig. 6.

Consequently the core of the transformer is not magnetically polarized by d.c., as it is in the half-wave system, and this is a point in favour of the full-wave system. If it were not for the reservoir C, which distorts the current waveform into sharp peaks, the current would be sinusoidal, and the transformer would be working under the most comfortable conditions possible. As it is, however, the current is distorted, as in Fig. 7. Now here is a question. The r.m.s. value (E) of the transformer voltage is known, and for a given set of rectifier conditions and hence waveform the r.m.s. value (I) of the transformer current can be calculated. The question is, what is the power supplied by the transformer? Is it EI? In ordinary a.c. theory it would be, provided that E and I were in phase. Well, here they do seem to be, as near as makes no matter. But if you start calculating the power used up in the load, and add in the power lost in the rectifiers and transformer wire, you soon find that the power delivered by the transformer can't be accounted for on this basis. So, as there is no possibility of deceiving Nature's auditor with regard to the power balance sheet, the basis must be wrong.

Watch the Waveform

Well, it is just another example of how theory limited to sine waveform lets one down if applied to other waveforms regardless. As I mentioned some years ago while discussing phase,† it is not really allowable to compare the phases of dissimilar waveforms. It is true that the fundamental component of the Fig. 7(b) current is practically in phase with the

† *Wireless World*, May and June, 1948.

voltage, but the current is certainly not all fundamental. The peakier its waveform, the greater is the proportion in the form of harmonics. And while some half-cycles of these harmonics are admittedly flowing in the same direction as the generator voltage, others flow against it so represent negative power (i.e., power flowing back into the generator), and tend to cancel out the positive power. Just what the net result is necessitates a not-so-easy integration; but without knowing any integral calculus at all we can make sure that when the current is very peaky the power is *not* equal to EI. Suppose that the current is I_{max} for one-tenth of each half-cycle, from 81° to 99°, and zero elsewhere. During this time the voltage varies from 0.988 E_{max} to E_{max} and back again. Its average is clearly above 99% of E_{max}, so we shall be near enough if we say it is E_{max}. The power during this period is therefore practically $E_{max}I_{max}$, and for the rest of the half-cycle is zero. The average power is therefore one-tenth of this—0.1 $E_{max}I_{max}$. Now compare this with EI. E is $E_{max}/\sqrt{2}$. While it is flowing, the current is I_{max}; the current-squared is I^2_{max}; the mean current-squared is $I^2_{max}/10$; the root mean current-squared or r.m.s. value is $I_{max}/\sqrt{10}$. So EI is $E_{max}I_{max}/\sqrt{20} = 0.223\ E_{max}I_{max}$. In other words, the actual power is not EI but

$$\frac{0.1\ EI}{0.223} = 0.447EI\ —$$

less than half what it would have been with a sine current waveform having the same r.m.s. value.

But although the r.m.s. value of the current, such as would be indicated by an accurate thermojunction milliammeter, doesn't necessarily count in reckoning power delivered, it counts only too well in reckoning heating of any resistance it passes through, so again it is the value that must be used for choosing the gauge of wire for the transformer secondary.

Summary

Let us sum up the voltage and current lore concerning the full-wave reservoir rectifier circuit. Voltages present little difficulty in practice: the "generator" (i.e. transformer secondary) usually supplies something reasonably sinusoidal, which can therefore be measured with any reasonably accurate voltmeter having an appropriate a.c. range. The significant voltage at the output is the mean value, measured with a moving-coil voltmeter. So far, the ordinary multi-range test meter is sufficient. The actual peak inverse voltage across either rectifier is important, but one doesn't usually bother to measure it, because the most it can be is a trifle less than twice the *peak* half-secondary voltage (or, if you prefer, a trifle less than the peak voltage across the whole secondary). The reason is pretty obvious: the greatest instantaneous voltage the generator can apply in the "wrong" direction is E_{max}, and at that moment C is also applying a voltage in the same direction, which at no-load would be very nearly E_{max}. So to be on the safe side the peak inverse voltage with this circuit is always taken as $2E_{max}$, or $2\sqrt{2}E = 2.8E$.

If one is interested in ripple voltage, then certainly complications do arise, because it has a decidedly non-sinusoidal waveform. But usually there is no need to distinguish very clearly between the several possible values—a rough idea will do. The peak value can be read by means of a peak valve voltmeter with blocking capacitor, or alternatively an oscilloscope.

Currents are trickier. There are:

(1) The mean value of the output current, measured by a moving-coil meter in series with R_L. This should equal the sum of the readings on the same meter connected to measure each rectifier current in turn.

(2) The peak value of the current through each rectifier. This is important for voltage rating. Use about 5Ω of resistance and a sensitive peak voltmeter or oscilloscope.

(3) The r.m.s. value of current through each rectifier. This is measurable with a thermal meter and is the one that must be used for calculating watts loss.

(4) The equivalent r.m.s. value of sinusoidal current in phase with the voltage. Presumably this could be measured by means of a suitable wattmeter, by dividing the watts supplied from the whole secondary by the r.m.s. voltage of the same. Lacking the suitable wattmeter (as most people do), one would have to reckon up the wattage bit by bit: chiefly the load power (load current × load voltage), plus twice the square of the r.m.s. current through either rectifier multiplied by the effective resistance of either rectifier and any limiting resistance in series with it.

(5) The ripple current through C. This is necessary for checking that the rating for the capacitor is not being exceeded. If a r.m.s. current meter is available, it can be read directly by connecting the meter in series with C. Because the ripple voltage waveform is far from sinusoidal—and the current waveform still farther from it, because the reduced reactance of C at higher frequencies favours the harmonics—the reading on a rectifier-type meter is likely to be somewhat out, but perhaps good enough for the purpose.

A disadvantage of the reservoir type of rectifier circuit, half-wave or full-wave, is that if the load is absent—as happens if the rectifiers come into action before the valves in the set have warmed up—the output voltage rises to nearly E_{max}. So if the smoothing capacitors have been chosen on a $E_{max}/\sqrt{2}$ basis they are not likely to be very happy during these periods. For this and other reasons (such as avoiding high rectifier peak current) use is sometimes made of the "choke input" circuit, in which a so-called swinging choke is connected between the rectifiers and C. This type of circuit works quite differently from any we have discussed, and would demand several pages to itself, moreover it is practically never used in domestic equipment, so I am only just mentioning it to show that when you have explored all the voltage-doubling etc. types there are still more worlds to conquer!

Receiving Valve Manual

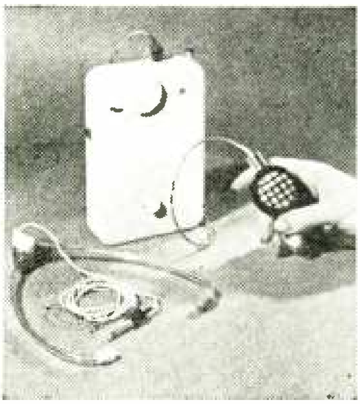
VALVES for colour television are included in the latest edition (RC-17) of the well-known RCA "Receiving Tube Manual" which is now available in this country. The book has been revised, expanded and brought up to date, and contains technical data on more than 500 valves and c.r. tubes. There are also sections on basic theory, interpretation of data, applications, installation and testing which contain a good deal of new matter. The section on circuits now includes several new circuits for use in high-quality audio amplifiers. Among these are a low-distortion input amplifier stage, a two-stage input amplifier using cathode-follower output, a bass and treble tone-control amplifier stage and a complete 10-watt high-quality amplifier.

Priced at sixty cents in the U.S.A., the manual is available in this country from RCA Photophone, Ltd., 36, Woodstock Grove, Shepherds Bush, London, W.12, at 8s.

Pocket Wire Recorder

PLAYING times of up to $2\frac{1}{2}$ hours are provided in the German-designed "Minifon 54" recorder which is to be distributed in this country by the "Emidicta" Division of E.M.I. Sales and Service, Ltd., 363, Oxford Street, London.

Measuring $6\frac{5}{8} \times 4\frac{3}{8} \times 1\frac{3}{8}$ in and weighing only $2\frac{1}{2}$ lb with batteries, the "Minifon" is a remarkable example of miniaturization, particularly in the mechanical drive mechanism and controls, which are of watch-like precision. A 12-V layer-built battery costing 5s 6d gives a running time of 10 to 15 hours. The motor, which is governed, drives a large diameter take-up spool which ensures a virtually constant winding speed of 11.8 in (30 cm)/sec: re-winding is at $2\frac{1}{2}$ times this speed. The wire diameter is 0.002 in.



"Minifon 54" wire recorder with lapel microphone and stethophone earpiece.

A three-stage amplifier using hearing-aid type valves is used for recording and playback, and runs from separate 1.5-V and 30-V batteries. The recorded frequency range claimed is 200-4,000 c/s and the quality of reproduction with saturation bias (a permanent magnet is used for erasure) is more than adequate for speech.

The basic price of the "Minifon 54" is £85 with batteries, 1-hour duration spool, lapel microphone, stethophone earpiece, and leather case. Numerous accessories are available including a wrist microphone, typists' foot-control unit, mains power supply unit for the motor, and a pick-up coil for recording two-way telephone conversations.

News from the Clubs

Birkenhead.—The Wirral Amateur Radio Society continues to meet at 7.30 on the first and third Wednesdays of each month at the Y.M.C.A., Whetstone Lane, Birkenhead. Sec.: A. C. Wattleworth, 17, Iris Avenue, Claughton, Birkenhead, Cheshire.

Birmingham.—"Television Aerials" is the subject of the talk to be given by A. P. Hale, of Belling & Lee, to members of the Slade Radio Society at their meeting on November 12th at 7.45 at the Church House, High Street, Erdington. The annual general meeting will be held on November 26th. Sec.: C. N. Smart, 110, Woolmore Road, Erdington, Birmingham, 23.

Portsmouth.—The Portsmouth & District Radio Society now has its own club room, open every evening, at the British Legion Club, Queen's Crescent, Southsea. Meetings are held each Tuesday at 7.30. The November programme includes films (2nd), a lecture on televison (9th) and a discussion on television interference (23rd). The club operates a QRP transmitter, G3DIT. Sec.: L. B. Rooms (G8BU), 51, Locksway Road, Milton, Portsmouth, Hants.

Cleckheaton.—At the November 3rd meeting of the Spen Valley & District Radio & Television Society D. Skirrow (G3GFD) will speak on "Radio Valves and their Uses." The subject for the meeting on November 17th is "Oscilloscopes" by G. F. Craven, of Craven Electronic Instruments. Meetings are held at 7.30 at the Temperance Hall, Cleckheaton. Sec.: N. Pride, 100, Raikes Lane, Birstall, Leeds, Yorks.

Newark.—Technical films will be shown at the meeting of the Newark & District Amateur Radio Society at 7.0 on November 7th at the Northern Hotel, Newark. At the mid-monthly meeting at Northgate House at 7.0 on November 18th a commercial trans-receiver will be demonstrated. Sec.: J. R. Clayton, 160, Wolsey Road, Newark, Notts.

Romford.—Meetings of the Romford & District Amateur Radio Society are held each Tuesday at 8.15 at R.A.F.A. House, 18, Carlton Road, Romford. Details of the winter programme, which includes lectures, discussions and films, are available from the secretary, N. Miller, 18, Mascalls Gardens, Brentwood, Essex.

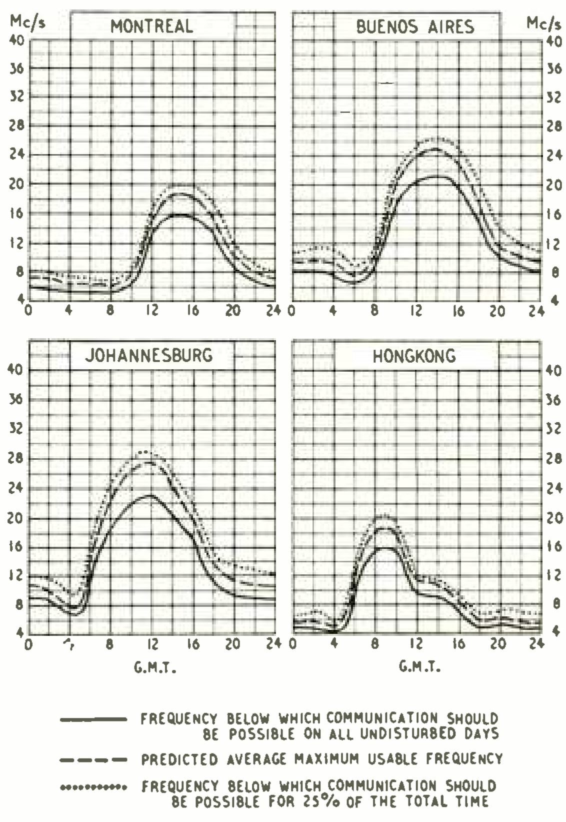
Southend.—Meetings of the Southend & District Radio Society are temporarily being held at the Ekco Works, Southend-on-Sea, on alternate Fridays at 7.45, the November meetings being on the 12th and 26th. Sec.: J. H. Barrance, M.B.E. (G3BUJ), 49, Swanage Road, Southend-on-Sea, Essex.

Short-wave Conditions

Predictions for November

THE full-line curves given here 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 November.

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



Measurement of Phase and Amplitude

Simple Method for Use with Feedback Amplifiers

By H. H. OGILVY, D.L.C.(Eng.), A.M.I.E.E.

IT is well known that high-gain negative feedback amplifiers tend to become, in fact, positive feedback amplifiers at extreme frequencies and may, unless special phase shifting and attenuating circuits are introduced, actually oscillate and become useless for the purpose intended. Feedback amplifiers, such as those used for computing are especially likely to be troublesome, since the feedback is usually about 100%. It is very desirable to be able to measure the characteristics of a prototype amplifier so that the tendency to oscillate, or rather, the stability margin, may be determined and, if insufficient, the necessary steps taken to improve this margin. The data required to assess the performance is usually presented in the form of a Nyquist diagram, and the preparation of this diagram requires the measurement of relative phase and amplitude of output with respect to input under open loop conditions over a wide range of frequencies. In particular, the measurement of phase presents some difficulty.

The conventional method is to use a cathode-ray oscilloscope and estimate phase from a Lissajous figure. This is rather clumsy and uncertain since the characteristics of the C.R.O. amplifier cause phase shift, particularly at the lower and upper frequencies. Commercial equipment is available for the measurement of phase, but is usually rather costly and complex. The method to be described is simple and economical.

If two sine waves of the same amplitude and frequency but differing in phase by an angle ϕ are added together, the resultant is also a sine wave, i.e., the resultant,

$$\begin{aligned} v &= V\sin\omega t + V\sin(\omega t + \phi) \\ &= V\sin\omega t + V\sin\omega t\,.\cos\phi + V\cos\omega t\sin\phi \\ &= V(1+\cos\phi)\sin\omega t + V\sin\phi\cos\omega t \\ &= V\sqrt{(1+\cos\phi)^2 + \sin^2\phi}\,\sin(\omega t+\alpha) \\ &\qquad \text{where } \alpha = \tan^{-1}\frac{\sin\phi}{1+\cos\phi} \\ &= V\sqrt{\cos^2\phi + \sin^2\phi + 1 + 2\cos\phi}\,\sin(\omega t + \alpha) \\ &= \sqrt{2}V\sqrt{1+\cos\phi}\,\sin(\omega t+\alpha) \end{aligned}$$

Hence the amplitude of the resultant is proportional to $\sqrt{1+\cos\phi}$ where ϕ is the phase angle. The deflection of a C.R.O. or a moving coil rectifier instrument would therefore be dependent on this function. The maximum value of $\sqrt{1+\cos\phi}$ occurs when $\phi = 0$ and if full-scale deflection (D = 1) on the indicating instrument occurs when $\phi = 0$ then $D = K\sqrt{1+\cos\phi}$, where K is a constant. Hence $1 = K\sqrt{1+1}$ and $K = 1/\sqrt{2}$ and therefore $D = \frac{1}{\sqrt{2}}\sqrt{1+\cos\phi}$ or $\cos\phi = 2D^2 - 1$ and $\phi = \cos^{-1}(2D^2-1)$.

Fig. 1 shows the curve relating phase angle to relative deflection (full line). Phase angles between 80° and 180° are easily determined from the deflection, but below 80° the rate of change of phase with deflection becomes too great for reasonable accuracy of

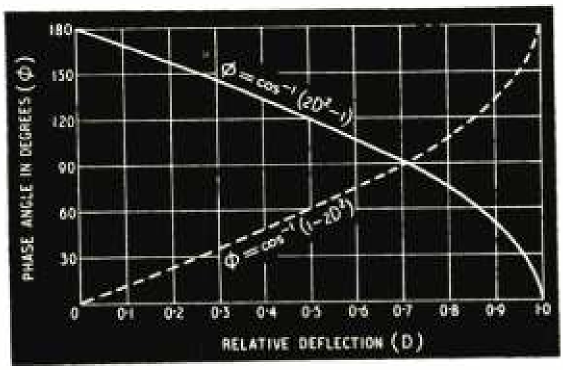


Fig. 1. Relative amplitudes resulting from the addition of two sine waves of equal amplitude and frequency, but with different phases.

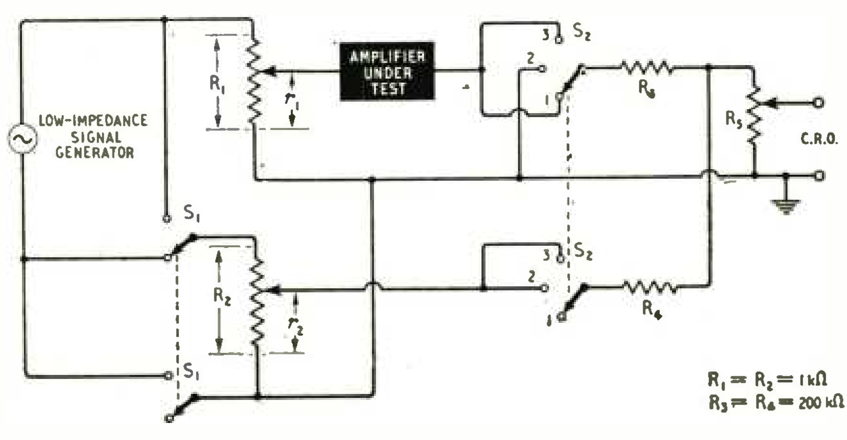


Fig. 2. Schematic diagram of phase measuring apparatus.

observation. However, if the reference signal, V sin ωt is reversed in phase the resultant wave is now

$$v = -\,V \sin \omega t + V \sin (\omega t + \phi)$$

and this gives a resultant amplitude of $\sqrt{2}V\sqrt{1 - \cos\phi}$. The relative deflection (D) is now 0 when $\phi = 0$ and 1 when $\phi = 180°$, if the constant of proportionality is obtained as before. Hence the expression $\phi = \cos^{-1}(1 - 2D^2)$ is obtained and this is also plotted in Fig. 1 (broken line) showing that phase angles between 0° and 100° may be accurately determined. It should be understood, of course, that this method does not differentiate between leading and lagging angles, but where there is a doubt the sense can be easily determined. In general, the condition required, i.e., that both signals to be compared must be of equal amplitude, will not apply but the method to be described shows how this may be achieved quite simply.

The basic scheme for measurement of phase and amplitude is shown in Fig. 2. It is assumed that a low-impedance signal generator is fed into two potentiometers of equal and relatively low resistance (about 1,000 ohms). With the switch, S_1, in the position shown, equal antiphase voltages are available at the wipers of R_1 and R_2 with respect to earth. This condition is therefore suitable for the measurement of phase angles between 0° and 100°. The amplifier under examination is fed from the wiper of R_1 which is set to some convenient value ($= r_1$) sufficiently small to avoid saturation of the amplifier. With S_2 in position 1 the deflection of the C.R.O. is adjusted, using R_1, to 50% of a predetermined arbitrary magnitude. S_2 is then placed in position 2 and R_2 adjusted for the same deflection ($= r_2$). The condition of equality of amplitudes has thus been achieved and the ratio $r_2\,r_1$ is obviously the gain of the amplifier at the particular frequency.

If S_2 is now placed in position 3, the output of the amplifier $=$ V sin $(\omega t + \phi)$ is now added to the reference $= -$ V sin ωt in the network R_3, R_4 and R_5. The arbitrary full scale deflection will occur when $\phi = 180°$. The relative deflection obtained will be according to the law $\phi = \cos^{-1}(1 - 2D^2)$ and ϕ may be determined from the curve in Fig. 1.

For phase angles greater than 100°, S_1 should be placed in the upper position and then $\phi = \cos^{-1}(2D^2 - 1)$. In the majority of cases, the sense of the phase angle will not be in doubt. Where there is uncertainty, however, the sense may be determined by making the output of the amplifier lag, using a simple RC circuit and repeating the above procedure. If the angle obtained has increased, then the original angle must have been negative or lagging. The facilitate measurements, the C.R.O. time base should be switched off, except when checking the amplifier output for saturation.

Although the scheme shown in Fig. 2 will be satisfactory for many purposes, there may be errors in measurement at the higher frequencies due to unequal impedances to earth at the terminals of the signal generator. Also, in some applications, it may be undesirable to load the device under test, even though the load (R_3) is several hundred thousand ohms. To overcome these difficulties, the circuit shown in Fig. 3 is suggested. This presents a very high impedance to the amplifier. Phase splitting is obtained electronically by V1a which is one-half of a double triode (12AT7). The measurement procedure is the same as before. The potentiometer R_5, may be fed into an amplifier of low output impedance and used to drive a moving coil rectifier meter. For convenience this meter may be scaled according to the law relating phase angle (ϕ) to relative deflection (D) (Fig. 1) and phase read off directly.

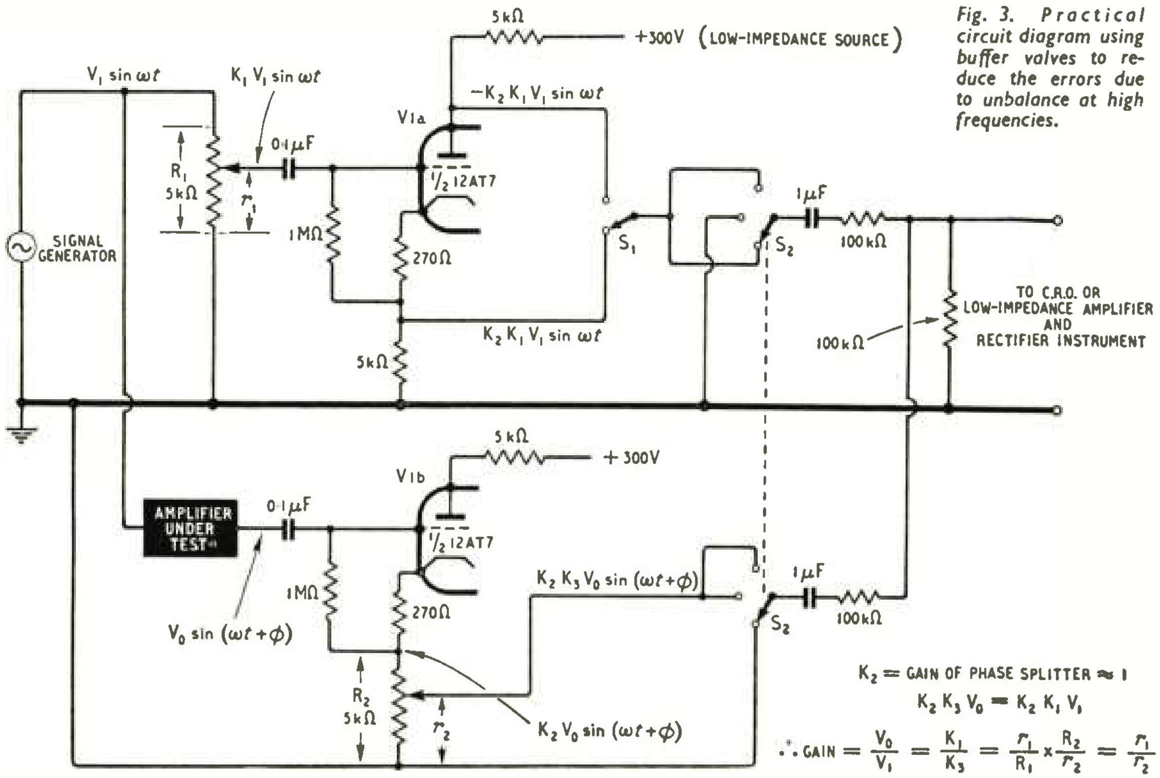


Fig. 3. Practical circuit diagram using buffer valves to reduce the errors due to unbalance at high frequencies.

NOVEMBER MEETINGS

Institution of Electrical Engineers

London.—November 8th. Discussion on "Methods of Teaching Technical Writing" opened by G. Parr at 6.0.

November 10th. "Standard Frequency Transmissions" by L. Essen, D.Sc., Ph.D., at 5.30, followed by "The Standard Frequency Monitor at the National Physical Laboratory" by J. McA. Steele, B.Sc.(Eng.), and "Standard Frequency Transmission Equipment at Rugby Radio Station" by H. B. Law, B.Sc.Tech.

November 16th. Celebration of the Jubilee of the Thermionic Valve commencing at 2.30 with the address of the Lord President of the Council, the Marquess of Salisbury, followed by "The Genesis of the Thermionic Valve" by Professor G. W. O. Howe, D.Sc., LL.D. At 3.30 "Thermionic Devices from the Development of the Triode up to 1939" by Sir Edward Appleton, D.Sc., LL.D., F.R.S., and at 5.30 "Developments in Thermionic Devices since 1939" by J. Thomson, M.A., Ph.D., D.Sc.

November 22nd. "Plastics for the Radio Engineer" by Maldwyn Jones at 5.30.

November 23rd. "The Application of the Hall Effect in a Semi-Conductor to the Measurement of Power in an Electromagnetic Field" by Professor H. E. M. Barlow, Ph.D., B.Sc.(Eng.), at 5.30 followed by "Audio-Frequency Power Measurements by Dynamometer Wattmeters" by A. H. M. Arnold, Ph.D., D.Eng.

November 30th. Discussion on "The Servicing of Electronic Measuring Instruments and its Effect on their Design" opened by Denis Taylor, M.Sc., Ph.D., at 5.30.

All the above meetings will be held at Savoy Place, London, W.C.2.

East Midland Centre.—November 9th. "Properties and Application of High Permeability Magnetic Alloys" by G. A. V. Sowter, Ph.D., B.Sc., at 6.30 at Loughborough College.

November 23rd. "Telemetering for System Operation" by R. H. Dunn. B.Sc., and C. H. Chambers, at 6.30 at the Gas Dept., Demonstration Theatre, Nottingham.

November 26th. "A Radio Position Fixing System for Ships and Aircraft" by C. Powell at 6.30 at the College of Technology, Leicester.

North Midland Centre.—November 9th. Discussion on "The New I.E.E. Examination Regulations" opened by E. C. Walton, B.Eng., Ph.D., at 6.30 at 1, Whitehall Road, Leeds.

North-Western Radio Group.—November 24th. "The Manchester-Kirk o'Shotts Television Radio-Relay System" by G. Dawson, B.Sc., L. L. Hall, K. G. Hodgson, B.A., R. A. Meers, and J. H. H Merriman, M.Sc., at 6.45 at the Telephone House, Chapel Street, Salford.

South Midland Centre.—November 22nd. "Loudspeaker Systems—Recent Trends in Design" by Major A. E. Falkus, B.Sc.(Eng.), at 6.0 at the James Watt Memorial Institute, Great Charles Street, Birmingham.

November 25th. "Colour Television" by C. J. Hirsch at 7.15 at the Winter Gardens Restaurant, Gt. Malvern.

Reading District.—November 29th. "Television Interference" by K. R. Seamans at 7.15 at the George Hotel, Reading.

British Institution of Radio Engineers

London Section.—November 24th. "The Development and Design of Direct-Coupled Oscilloscopes for Industry and Research" by M. J. Goddard at 6.30 at the London School of Hygiene and Tropical Medicine, Keppel Street, W.C.1.

Scottish Section.—November 4th. "The Latest Developments in TV Cameras" by H. McGhee at 7.0 at the Institution of Engineers and Shipbuilders, Elmbank Crescent, Glasgow.

Merseyside Section.—November 4th. "Radio Receiving Valve Manufacture" by G. P. Thwaites, B.Sc., at 7.15 at the College of Technology, Byrom Street, Liverpool, 3.

North-Western Section. — November 4th. "Electronic Servo Mechanisms" by J. L. Russell at 7.0.

November 30th. "Electronics and the Wind Tunnel" by G. J. Scoles, B.Sc., at 7.0.

Both meetings will take place at the Reynolds Hall, College of Technology, Sackville Street, Manchester.

North-Eastern Section. — November 10th. "Stereophonic Sound" by R. A. Bull, B.Sc.(Eng.), at 6.0 at Neville Hall, Westgate Road, Newcastle-upon-Tyne.

South Wales Section.—November 17th. "The Techniques of Power Measurements from D.C. to 5 Mc/s" by G. F. Lawrence at 6.30 at the College of Technology, Cathays Park, Cardiff.

British Sound Recording Association

London.—November 19th. "Balance and Control" by G. Elliott at 7.0 at the Royal Society of Arts, John Adam Street, W.C.2.

Manchester Centre.—November 22nd. "Transformers and Chokes" by J. S. Holiday at 7.30 at the Engineers' Club, Albert Square, Manchester.

South-Western Centre.—November 24th. "Hi-Fi Can Be Music" by N. C. Mordaunt (Tannoy) at 7.30 at Callard's Café, Torquay, Devon. (Joint meeting with Incorporated Practical Radio Engineers).

Physical Society

November 5th. Duddell Lecture, "The Development and use of Large Radio Telescopes," by Professor A. C. B. Lovell at 5.0 at Burlington House, London, W.1.

Television Society

London.—November 12th. "Faulty Interlacing" by G. N. Patchett, Ph D., B.Sc., at 7.0 at the Cinematograph Exhibitors' Association, 164, Shaftesbury Avenue, W.C.2.

November 18th. Conversazione to mark the Jubilee of the Invention of the Thermionic Valve at 7.0 at University College, Gower Street, W.C.2.

November 25th. "European Television Programme Exchanges" by M. J. L. Pulling, O.B.E., M.A., (B.B.C.) at 7.0 at the C.E.A., 164, Shaftesbury Avenue, W.C.2.

British Kinematograph Society

London.—November 24th. "Factors Affecting Quality in Colour Television" by I. J. P. James, B.Sc., at 7.15 at the Gaumont-British Theatre, Film House, Wardour Street, W.1.

Radio Society of Great Britain

November 19th. Technical films at 6.30 at the I.E.E., Savoy Place, London, W.C.2.

RANDOM RADIATIONS

By "DIALLIST"

TV Interference with Radio

THERE IS perhaps far more interference radiated by television receivers, to which J. Platts and G. O. Thacker refer in last month's *Wireless World,* than is generally realized. Many owners of broadcast receivers have come to regard any unwanted noise from the loudspeaker as just one of those things and don't know that, if they will only report it, the P.O. engineers will do their best to help. No one should be allowed to operate a television receiver which spreads alarm and despondency among his listening and viewing neighbours. But the P.O. anti-interference people either don't realize the existence of clause No. 4 of the television licence schedule or have been instructed not to enforce it—or to prohibit the use of the offending set (under clause 7) until it has been rendered innocuous.

A Bit Much?

The people really to blame, of course, when television interference is broadcast, are not the users but the manufacturers. It never occurs to the ordinary completely non-technical buyer of a TV receiver that the set of his choice may possibly cause various unpleasant things to happen to his neighbours' receivers whenever he switches it on. The advertisements assure him that it is the best of the lot; his wife likes the cabinet; the 17-inch tube will take the wind out of the sails of the Robinson's 15-inch next door. He puts down his money, or signs a "never-never" agreement, and feels that a good job has been well done. Can't you imagine his indignation and his "Pygmalion" retort when the P.O. engineers diffidently suggest that he should have something done to his beautiful set and pay for it? A warning to manufacturers by the P.M.G. that after a certain not-too-far ahead date a ban on the use of interfering receivers would be enforced might work wonders. The new B.R.E.M.A. standards for manufacturers should bring about a great improvement; but not all makers of television receivers are members of the association.

Anti-Flutter

WITH the rapid expansion of both civil and military aviation each year sees more and more planes in the air. Unless we live in remote places they now pass over or near our homes at all hours of the day and night. Aeroplane flutter is already one of the commonest kinds of interference with television, particularly in places near aerodromes, or on regular flying routes. No form of a.g.c. seems able to cope adequately with flutter, for the time constants of the circuits are too long. Here's an opportunity for someone to develop an effective system of "automatic flutter suppression." It's sure to be done some day. As a dweller within a couple of miles of a big and busy aerodrome, I hope fervently that it may come soon.

All-Dry Sets

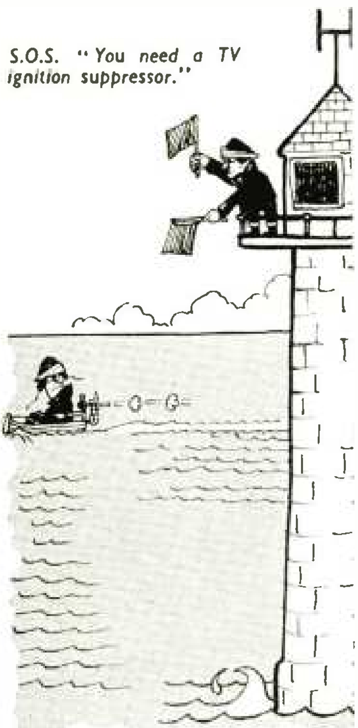
THE "ALL-DRY" receiver has many attractive points; and there is still more to be said for the kind that can be worked from the mains or from its own batteries. The "pros" for both sorts are that they are light, genuinely portable and of small size. But there is one serious "con" and that is the comparatively short life of the rather expensive combined h.t. and l.t. battery. Actually, it's the filament part of this battery which passes out, as a rule, and leads to distortion and, eventually, to a signal too weak to be of any use. When the filament section has packed up the h.t. cells would usually be good for many more hours of useful service. Would it not be a great improvement if the h.t. and l.t. dry batteries were separate units? I haven't a doubt that it would; nor should there be any difficulty about it for set makers or battery manufacturers.

Time by the Forelock

DURING a recent visit to East Anglia I noticed over one house the familiar vertical H television aerial and a very much smaller horizontal array. The orientation appeared to be about right for Wrotham; but could this moderately powered f.m. transmitter possibly be receivable at so great a distance? The answer came a day or two later when I happened to meet the owner of the house. "Do you find Wrotham any good here?" I asked. "Wrotham," he said; "What's that?" "Why, the B.B.C. experimental f.m. station. Isn't that what your horizontal array's for?" He laughed: "Oh no," he said, "that's for Norwich, when it gets going." Nothing like being in good time! The temporary Norwich TV station, using, I believe, the present channel 3 Brighton "booster" as its transmitter, isn't due to make a start for 5 or 6 months yet.

Modern Mains Receivers

THE thing that I like least about our mains receivers of to-day is that the great majority are built on a.c./d.c. lines with no transformer between them and the supply mains. On d.c. there's nothing much amiss, for neither sound nor television receivers will work unless the mains plug is put the right way round into its socket. But on a.c. both will function whichever way it's inserted. According to the law of averages this means that at any time half the mains sets operated on a.c.—say 5,000,000 "sound" and 1,500,000 television—are working with their chassis directly connected to the phase wire and with nothing earthed. The thought of all those acres of chassis swinging through 500-700 volts peak-to-peak 50 times each second must be a rather solemn one to any electrical engineer brought up on the old sound principles. Ours is, I believe, the only civilized country in which this sort of thing is permitted and I can't help feeling sorry that the I.E.E. regulation about the use of an isolating transformer between a.c. mains and apparatus was ever relaxed. But there it is, and we've got to make the best of it. As a safeguard, my suggestion is that all mains leads should be fitted with sockets which can be connected in one way only to the plugs in the receiver and that at the other end there should be a 3-pin plug. Those who install receivers should be required to see that the a.c. power supply is from a correctly connected 3-pin socket.



S.O.S. "You need a TV ignition suppressor."

UNBIASED

By FREE GRID

Stars or Sunspots?

FORECASTING the future by means of the calculated positions of the heavenly bodies at certain given times is known as astrology when carried out by Madame Estelle or by those journals which cater for the unlettered masses, and astronomy when employed by the Astronomer Royal to give us the dates and times of tides and eclipses.

Now, astrologers usually confine their prophecies to things like love, marriage and other disasters which depend on the whimsy of women and it is not very surprising therefore, that their percentage of success is so low. The Astronomer Royal on the other hand, eschewing women and their ways, bases his forecasts on the solid rock of science and so obtains 100% success.

I draw attention to these facts because something quite new in the realm of these stellar forecasts has made its appearance and this something concerns we radio men very much indeed. A super-modern stargazer hailing from the U.S.A. claims to be able to forecast ionospheric conditions for radio transmissions at certain times by studying the stars instead of sunspots.

This celestial observer has secured the backing of one of the largest radio organizations in America, which has published figures showing that his successful forecasts represent 92% of the whole. However, in the opinion of some British ionospherists, this rather startling figure is obtained by a method of relating forecasts to ionospheric disturbances which, technically speaking, is open to question.

I am inclined to agree with this opinion, although I think we ought to give this planetary pundit more rope with which to hang either himself or us. Any good racecourse tipster can produce first-class results over a limited period but he cannot keep it up. Indeed, we had a remarkable instance of this sort of thing in the middle of the war when the editor published a letter under the heading of "What the Sunspots Foretell."* The writer of this letter pointed out a striking correlation between the sunspot cycle and the career of Hitler and then proceeded to use this correlation to prophesy the end of the war with some degree of accuracy.

"History is Bunk"

WHY is it that the Radio Industry Council took such pains to stress that this year's radio show was the 21st when actually it was the 25th? I believe that the official explanation is that the shows of the 1922-25 period were not "National" shows, and yet in a potted "Radio Show History" which the R.I.C. issued to the Press at the time of the show we are told that one of them was promoted by the *National* Association of Radio Manufacturers and Traders (NARMAT). What the R.I.C. means, I suppose, is that the pre-1926 shows were not organized by its progenitor—the Radio Manufacturers' Association—because it was not until that year that this new name—and a new constitution too—was adopted. It is as though the B.B.C. issued a history of British broadcasting and ignored the pre-1927 years because the service was then conducted by the British Broadcasting *Company* and not the British Broadcasting *Corporation*.

The "historian" picks out home-construction as the one thing worthy of mention at the 1924 show and completely ignores the fact that a valve with the then unheard-of amplification factor of 20 was shown. This was also the first show at which the superhet was seen; two very prominent firms exhibiting it.

This official "history" describes the superhet as one of the novelties of the 1927 show, whereas in actual fact it was rather in eclipse then. It did not start to stage a real comeback until over three years later, as I myself mentioned in *W.W.* for September 17th, 1930. The real novelty about which everyone was talking in 1927 was the screen-grid valve. And so I could go on.

Sackcloth and Ashes

I CERTAINLY put my foot in it when I said in the October issue that the smallest tape recorder available weighed 12 lb and was far from pocketable. I have had several letters pointing out to me that there is one on the market weighing only 2½ lb which is really pocket-size. As it is described elsewhere in this issue, I will say no more about it except to plead in my defence that at the time I wrote I don't think this instrument was available in this country; I won't split hairs by claiming that I said tape recorder, whereas this instrument uses wire.

Electronic Pulse Taking

LAST JUNE I suggested that the present method of pulse-taking in hospitals was out of date, likely to be very misleading and a gross waste of the nurses' time. I pointed out that this could be done electronically and automatically. By means of a miniaturized v.h.f. transmitter strapped to each patient's wrist and a receiver with a battery of c.r. tubes in the ward-sister's room, heartbeats could be read at any moment and, if desired, recordings could be made on film or magnetic tape.

The Palpatron

Now I learn from the leading journal of the nursing profession (*Nursing Mirror*, August 27th) that an electronic device called the Palpatron is in use for this purpose at the Boston City Hospital, Massachusetts, U.S.A., and that its great sensitivity enables a pulse reading to be detected when nothing can be discerned by the normal manual method.

Τηλεόρασις

THROUGH the courtesy of a reader I learn that the terrible word τηλεβισιον is not used in modern Athens for television. No doubt some of you noticed that there was an unfortunate typographical error in my note last month, the omicron being omitted. The modern Greek word is τηλεόρασις but I don't think we should have taken kindly to teleorasis instead of television. The same reader tells me of an excellent and up-to-date telecommunications dictionary of English and Modern Greek, published by the Sitivanideos Institution in Athens, with funds provided by the American Mission (European Co-operation Administration) in Greece.

* The letter mentioned by Free Grid was published in our March 1942 issue.—Ed.