Wireless World

ELECTRONICS, RADIO, TELEVISION

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In This Issue

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TRANSISTORS IN AUDIO AMPLIFIERS (Part Two)

Although in principle a large number of circuits can be obtained by combining grounded emitter, grounded base or grounded collector configurations with transformer or R-C coupling, in practice transistor audio amplifiers tend to follow a simple pattern. A typical circuit can be considered to have grounded emitter stages in cascade, with R-C coupling, and with d.c. stabilisation provided by the potential divider and emitter resistor method.

The maximum power gain available with perfect matching (and transformer coupling) when the effective load resistance in the collector circuit $R_L = \sqrt{r'_{22} \cdot r'_{out}}$ and the effective source resistance $R_s = \sqrt{r'_{11} \cdot r'_{in}}$ is

$$\left(\frac{a'}{\sqrt{r'_{11}} + \sqrt{r'_{in}}}\right)^2 \cdot r'_{22}.$$

R-C coupling is preferred generally to transformer coupling for low cost and phase shift and good response, but the power gain of each stage then arises solely from the inherently high current gain of the grounded emitter stage, and the higher gain which would be available by impedance matching with the transformer is not achieved.

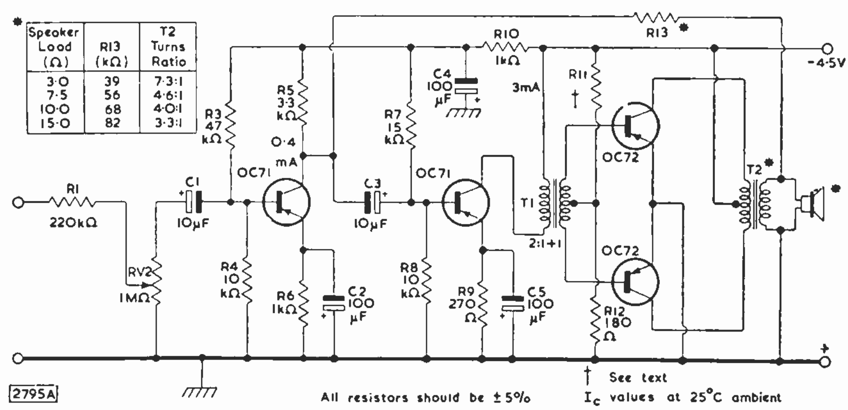
The factors entering into the design of an R-C coupled transistor cascade are not difficult to appreciate; many of them are similar to those encountered when working with valves. The collector voltage and current are limited by d.c. ratings $V_c$max and $I_c$max, and by a.c. ratings $v_{c(pk)}$max and $i_{c(pk)}$max. For high gain and output power the battery voltage should be high, but a lower voltage and hence smaller current drain is more economical. The high value of collector load resistance required for maximum gain cannot be obtained with R-C coupling, as there is no advantage in making the collector load very much greater than the effective parallel input impedance of the next stage. In addition, the load resistance and collector current determine the voltage available across the transistor, which is also reduced by the emitter resistance included for stabilising. The collector current should therefore be small so that a large collector load resistance can be used; on the other hand a large collector current swamps the variation in collector leakage current $I'_{c(o)}$ with temperature.

After allowing for these various conflicting claims, the number of stages is chosen to give the required overall gain when feedback is applied. Since the signal swing in the early stages is small, the d.c. working point can be chosen for low current drain (and noise), provided they have potential divider and emitter resistor d.c. stabilisation. The power gain in the grounded emitter R-C coupled stage can be calculated from $(a')^2 R_L/r'_{in}$, the a.c. current gain being a' and the voltage gain $a' R_L/r'_{in}$. This expression assumes that R_L is very much smaller than r'_{22} and r'_{out}.

Here, a', r'_{in}, etc. are Small-Signal parameters given in published data and computed for the working point employed. As the load on an R-C coupled stage is formed by its collector resistance in parallel with the input resistance of the following stage, the power and voltage gain for each stage can be calculated by working backwards through the cascade.

Class AB push-pull operation in which the bias corresponds very nearly to that for true Class B operation is a natural choice for the output stage when a transistor amplifier is to be designed as a power amplifier, that is, to give the highest output power permitted by the collector dissipation p_{cmax}, without objectionable distortion. The quiescent power consumption is very small and the efficiency is high. The Mullard OC72 is intended for this mode of operation. An actual circuit is shown in the diagram, the output power being 200mW for 10% total harmonic distortion for an input of about 6mV at C1 or 500mV at R1. Negative feedback is applied over the driver and output stages by R13, which is matched to the loudspeaker. A small amount of bias is provided to the OC72's by the potential divider R11–R12, which is effective in reducing the high crossover distortion inherent in a true Class B transistor output stage.

Speaker Load (Ω)	R13 (kΩ)	T2 Turns Ratio
3·0	39	7·3:1
7·5	56	4·6:1
10·0	68	4·0:1
15·0	82	3·3:1



† See text
All resistors should be ±5%
I_c values at 25°C ambient

The value of R11 must be chosen from the range 6.8, 6.2, 5.6, 5.1, 4.7, and 4.3kΩ so as to adjust the total quiescent current in the output stage to 1.3mA±10% at 20°C or 1.6mA±10% at 25°C. The operating ranges with speech and music are 15°C to 45°C ambient temperature and 4.5V to 2.7V (or even 2.0V, depending on the distortion tolerated by the listener) with a Leclanché type battery.

Suitable transformers can be obtained from R. F. Gilson Ltd. The phase splitter transformer is type WO780 and the output transformer WO781. The secondary resistance must be specified as 3.75Ω, 7.5Ω, or 15Ω when ordering the output transformer.

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Combined Audio Show?

WITHIN the space of six weeks, from April 13 to May 26, two shows covering the same subject and run on broadly similar lines were held in London. These were the Audio Fair and the annual exhibition of the British Sound Recording Association. That meant a duplication of effort on the part of most of the exhibitors (to say nothing of many visitors) which can hardly be allowed to carry over into future years. That was acknowledged by Norman Leevers, the retiring president of B.S.R.A., who, at the Association's annual dinner, speculated on possible ways of arriving at a more satisfactory arrangement.

Mr. Leevers put forward two alternative proposals: that the B.S.R.A. show should be moved to the autumn, or that it should be combined with the Audio Fair.

To us, the idea of a combined exhibition seems to be the better. Demonstrations under reasonably good conditions are an essential part of a sound reproduction show, and, practically speaking, the only suitable venue is a large hotel. In London, hotel accommodation is easier to obtain in the spring than in the autumn. And, anyway, it is doubtful whether the majority of manufacturers would continue to support two shows in the year.

An effective basis for collaboration between the organizers of the Audio Fair and the B.S.R.A. should not be too difficult to work out. Facilities might be provided for the B.S.R.A. to stage non-commercial exhibits and demonstrations of various aspects of sound reproduction. Then the Association's annual convention might be arranged to coincide with the Fair. Such things as these would, we feel sure, prove to be attractive to many visitors. Collaboration between the two bodies on these lines should be beneficial to both of them.

Recorded Programmes

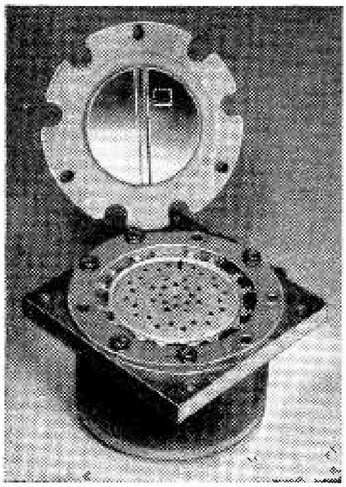
THOUGH the programme side of broadcasting is no real business of *Wireless World,* we sometimes feel impelled to comment on it when fundamental issues likely to affect the growth of the service are concerned. One such issue is that of transmissions from recordings; sometimes it seems to us there are too many of them, both on sound and (in the form of film) on television. Anyway, we cannot resist the temptation to quote from a leader in *The Times* (May 21). The newspaper's comments were linked to a recent lecture on the RCA system of television recording on magnetic tape, but could be applied with equal or even greater force to the use of films.

"One of the possibilities is that technical pressure may cause television to lose some of its spontaneity. There is nothing, except a determined will to do so, to stop the making of a television programme becoming something like that of a film. This could alter the whole artistic nature and scope of television. Again, there may come a strong tendency for the tail to wag the dog—the content of the programme being influenced by the fact that it is a recording. Some of these fears are theoretical. That does not mean to say they will not materialize."

Research and Measurement

THE number of new instruments and techniques shown at the Physical Society's exhibition, reported elsewhere in this issue, is still on the increase, and several new trends were noticeable. In particular, there was fairly widespread use of transistors in instruments, both as h.t. generators and, in some cases—e.g., in physiological work—where their small size and self-contained nature is an especial advantage. There was also a new type of transistor with quicker response than that of normal junction types. This device is perhaps symptomatic of the general development of junction transistors capable of working at higher frequencies than at present. The same tendency was evident in oscilloscope tubes with ability to record frequencies of the order of 1,000 Mc/s; these are specially adapted to the recording of very fast transients. On the same trend, super-high-frequency valves capable of being tuned over quite a wide range were shown.

A particularly commendable and much appreciated section of the exhibition was the demonstration of the principles of colour television, arranged by the Physical Society Colour Group.



Simple Wobbulator

Electro-Mechanical Modulation (6 Mc/s) in a Single-valve Circuit

By B. T. GILLING

Modulator unit with ceramic top plate removed, showing semi-circular silvered fixed electrodes and domed and perforated diaphragm.

THE wobbulator to be described reaches just about the limit of simplicity, consisting as it does of a single-valve oscillator; yet it is most satisfactory in use, and gives on Band I a linear sweep of six megacycles.

In the American radio altimeters now available on the surplus market there is a component which makes the above possible. The altimeters have the type numbers RT-40/APN, AN/APN-1, AN/ARN-1, AYB-1, AYD and there may be others. The component is often described in advertisements as a "magnetic sounder" but it would be more appropriately called an electro-magnetic frequency modulator.

A typical example is shown, partially dismantled, in the accompanying photograph. It consists of a magnet and coil, similar to that of a moving-coil loudspeaker, to which is attached a slightly domed aluminium diaphragm of about two inches diameter, freely suspended and perforated all over to prevent air loading. Mounted in front of this diaphragm on a ceramic cover are two metal plates and these with the diaphragm form a two-gang capacitor. The capacity swing of each section is from 10 to 50 pF. It will be obvious that if this capacitor is connected across the coil of an oscillator and the moving coil energized by an alternating current, the oscillator will be frequency modulated.

The circuit of the complete wobbulator is given in Fig. 1. Ample frequency deviation on Band I could be obtained by using only one section of the f.m. capacitor, but since it was also desired to sweep the i.f. band around 10 Mc/s the two sections were connected in parallel to give a maximum swing of 20-100 pF. This means that one side of the capacitor is earthed, making it necessary to use an electron-coupled type of oscillator. The 10-pF variable capacitor in parallel with the coil is brought out to a control on the panel to enable the middle frequency to be set.

Any miniature medium-mu triode is suitable for the valve, in fact one of the acorn triodes in the altimeter is an admirable choice. The frequency-modulated output is taken from the anode of this valve and its amplitude is controlled by a simple attenuator.

Alternating current for driving the moving coil is obtained from the heater supply. A series variable resistor controls its amplitude, hence the frequency deviation, and a fixed resistor prevents over-drive. The sweep voltage for the X-plates of the oscilloscope is also taken from the heater supply.

Return Trace Suppression

In an ideal case the forward and return traces would coincide exactly, but owing to phase shifts in the amplifiers this does not happen, and a trace as in Fig. 2(a) is always obtained. There are several ways of removing the second trace and the method adopted in the present case is to take a blanking voltage from a phase shift network across the heater supply and apply it to the grid of the tube. This not only blacks out the unwanted half of the trace but brightens the centre portion of the wanted half where its brilliance decreases owing to the increased speed of the spot. A single trace of even brightness throughout its length is obtained, as in Fig. 2(b).

At full amplitude of the moving coil a sweep of 10 Mc/s in Band I is easily obtained. This is illustrated in Fig. 2(c). With this amplitude, however, frequency linearity suffers, but for a sweep of 6 Mc/s it is very good, as will be seen from Fig. 3.

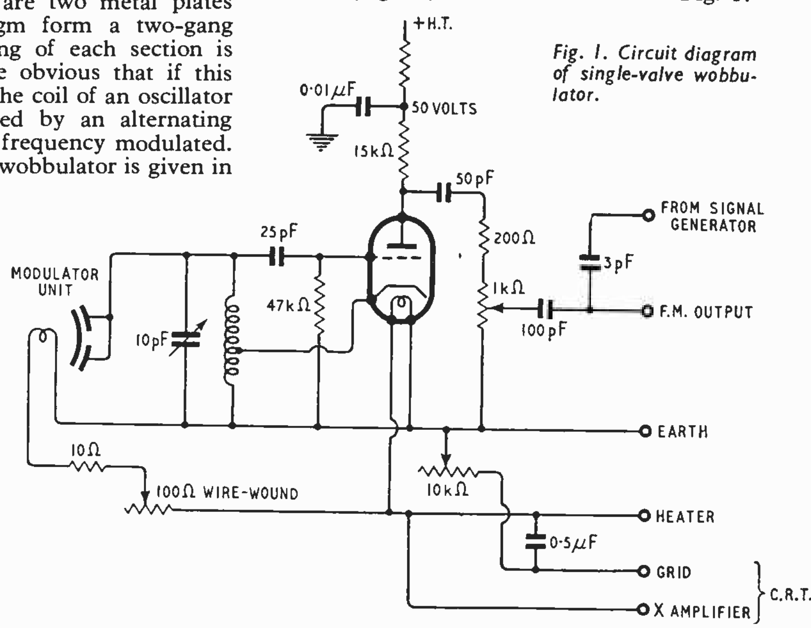


Fig. 1. Circuit diagram of single-valve wobbulator.

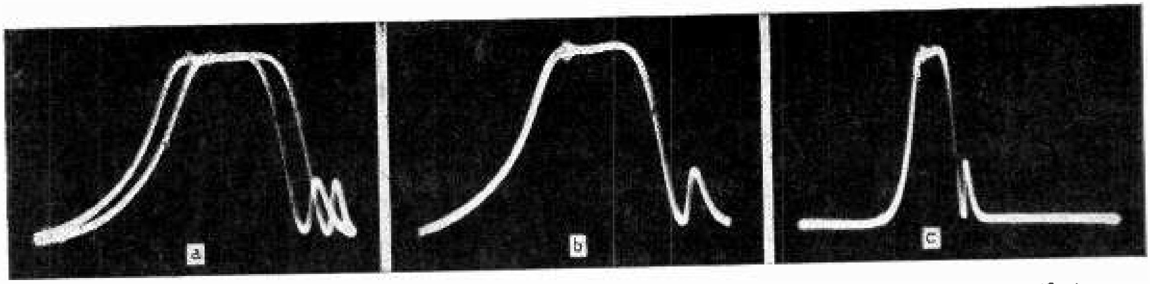


Fig. 2. Three forms of the same response from a partly aligned television receiver. (a) Displaced return half, due to phase shift. 6-Mc/s sweep. (b) Return half suppressed. 6-Mc/s sweep. (c) Same response with 10-Mc/s sweep.

The trace obtained is of little use unless some means are provided of ascertaining the frequency of all points along it, so a marked "blip" is applied by adding the output from a signal generator to the output of the wobbulator through a small capacitor. With a normal wide-range Y-amplifier this will appear as a thickening of the trace on each side of the marked frequency. To get a sharp blip the frequency range of the amplifier must be severely limited and the simplest way of doing this is to connect a 0.001-μF capacitor across its input. The resultant blip is seen in the three oscillograms in Fig. 2. As the frequency of the signal generator is varied the blip will travel along the trace marking the spot frequency at any given point.

This particular wobbulator was designed for Channel 3, Band I, and the i.f. band from 9 to 14 Mc/s with switched coils. Both coils are wound on half-inch formers and are as follows. Channel 3: 9 turns, tapped at 3 turns, spaced to occupy $\frac{3}{8}$ inch. I.F.: 18 turns, tapped at 6 turns, 34 s.w.g. close-wound. Adjustments to make the coils resonate at about the middle frequency can be made with either iron dust or brass cores.

Alternative Modulator

The heart of this wobbulator is the unit from the altimeter, but if this component is not obtainable it is possible to make a very satisfactory one from a three-inch moving coil loudspeaker. There are two ways of doing this, both of which entail the cementing of a thin aluminium disc to the diaphragm. In the first case a metal plate is fixed immediately in front of the disc as closely spaced as practicable to form the variable capacitor. In the other the actual oscillator coil is wound in pancake form and mounted closely to the disc. As the disc moves the changing eddy-currents alter the inductance of the coil. This latter method is not as satisfactory, as it prevents other coils being switched in to provide alternative ranges.

The only disadvantage of the generator in the form shown is that it requires a coil for each channel swept. At present, when there are only two, or at most three, channels operating in any area, this is of no real significance. To cover future development and the possibility of the introduction of colour television, which will call for the examination of a band almost in the audio region, this unit can be looked on as the basis of a more elaborate wobbulator of the beat frequency pattern. In this it would form the frequency-modulated oscillator which beats with a tunable oscillator to give a resultant swept output over a very wide range of middle frequencies.

This electro-mechanical method of obtaining frequency modulation is far simpler and more reliable than the reactor valve and has proved so satisfactory that it is used extensively, especially in America, for commercial wide-range beat frequency wobbulators.

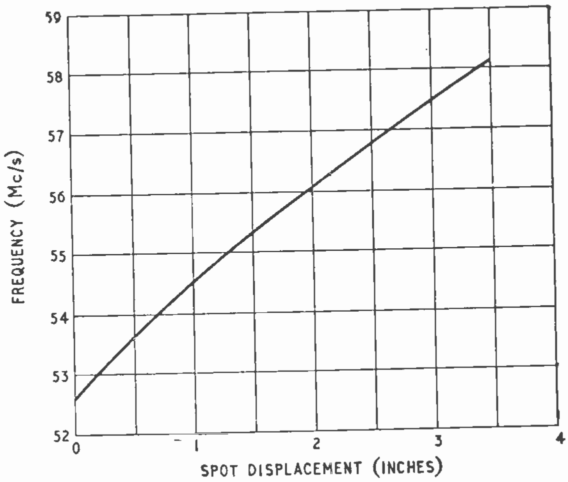


Fig. 3. Linearity of trace on Channel 3.

B.S.R.A. Convention and Exhibition

ONCE again the annual exhibition* and convention of the British Sound Recording Association was held in the Waldorf Hotel and the increased space made available was fully booked. Of the 42 firms exhibiting, 27 gave continuous demonstrations in separate rooms in the hotel, instead of sharing the use at specified times of a communal listening room. Conditions were more comfortable and there was much less fluctuation in the density of visitors in the main exhibition than on previous occasions.

Much interest was shown in the competition by members of the Association for the best amateur-constructed equipment. The winner of the President's Trophy was V. L'Estrange for a versatile lightweight magnetic recorder with three tape speeds. The *Wireless World* prize was awarded by the judges to G. A. Jeary for a synchronizing system for a film projector and magnetic tape recorder and the Committee Prize went to J. M. Beukers for a beautifully finished pre-tuned f.m. receiver and gramophone pre-amplifier.

At the annual dinner Norman Leevers, the retiring president introduced his successor, J. F. Doust (M.S.S. Recording Company) who has had a long association with technical developments in disc and tape recording.

* Sound reproducing equipment shown at this and other recent exhibitions will be reviewed in the July issue.—EDITOR.

WORLD OF WIRELESS

Organizational, Personal and Industrial Notes and News

National Radio Show

OF the eighty or so British radio equipment manufacturers included in the initial list of exhibitors at the National Radio Show (Earls Court, August 22nd to September 1st), about 50 per cent are makers of domestic sound and television receivers—the remainder being component and accessory manufacturers. With the addition of the "user" exhibitors, publishers, banks and wholesalers, the total number of stand-holders is 111.

It is anticipated that some 400 television receivers will be in operation at the Show where both Band I and Band III programmes will be distributed.

As in past years, the main part of the ground floor will be devoted to manufacturers' stands. On the first floor will be the displays of the B.B.C. and, for the first time, the I.T.A. and the London programme contractors. There will again be a "careers and electronics" display on this floor.

A brochure for prospective overseas visitors has been circulated and copies may be obtained at the various offices of the British Information Service.

Southern England TV

THE permanent aerial on the new 500ft mast at the B.B.C. television station at Rowridge, Isle of Wight, is being brought into service on June 11th. This will increase the e.r.p. to more than three times that provided by the temporary aerial system on the 200ft tower which has been in use since the station opened in November, 1954. As the new aerial is directional the vision e.r.p. varies from 1 to 32 kW. Provision is made for the mast to be used for v.h.f. broadcasting also.

The temporary transmitter on Truleigh Hill, near Brighton, will continue in service for the time being although it was originally intended only as a temporary measure until Rowridge was operating on full power.

Institution of Electronics Engineers?

ALTHOUGH originally formed in 1946 as something of an "old comrades' association" the Radar Association has recently taken on the aspect of a technical institution. As such it has arranged a number of technical meetings during the past year at which lectures on such topics as underwater television, guided missiles, radio astronomy and colour television have been given. It is now planned to start a students' section for the benefit of those studying electronics.

Sir Robert Renwick, the president, speaking at the 10th anniversary dinner of the Association, referred to the need for the electronics industry to have its own technical institution. "I believe," he said, "that the Radar Association is destined to assume this rôle." He suggested that the scope of the Association should be extended and that its name should be changed to cover the whole field of electronic engineering.

Because the final stages of the passages of the British Railways vessels on the Heysham-Belfast service have to be navigated stern first, the radar installation for the three latest vessels has been specially adapted by Kelvin Hughes. As will be seen in this photograph of the experimental installation at the K. & H. research station on Southend pier, the masts will be fitted with deflector plates to eliminate spurious echoes. The vessels, which incidentally are fitted with bow rudders, will have a second radar display unit on the after navigating bridge.

Birthday Honours

Sir Gordon Radley, C.B.E., who has been director general of the Post Office since 1955 and was previously engineer-in-chief, is appointed a Knight Commander of the Order of the Bath.

John Anderson, C.B.E., chief scientist at the Admiralty Signal and Radar Establishment, Cosham, Hants, is appointed a Companion of the Order of the Bath.

L. H. Bedford, O.B.E., M.A., B.Sc.(Eng.), chief engineer, Guided Weapons Division of the English Electric Company, which he joined in 1947, is promoted to Commander of the Order of the British Empire. He was for many years director of research at Cossors and was one of the first two industrial engineers to be taken into the confidence of the Government on radar. He evolved the "Bedford" attachment for early gun-laying radar.

T. Constantine, chairman of Bonochord Limited, is appointed a C.B.E. for public and political services.

Appointments as Officers of the Order of the British Empire are conferred on **A. B. Howe**, M.Sc., A.R.C.S., assistant head of the research department of the B.B.C., which he joined in 1924, and **H. G. Sturgeon**, director and chief engineer of Ultra Electric.

J. Treadgold, B.E.M., lately principal station radio officer, Admiralty Civilian Shore Wireless Service, and **E. E. Frewin**, chief technical superintendent of the Gold Coast broadcasting department, are appointed M.B.E.

PERSONALITIES

Professor H. E. M. Barlow, B.Sc.(Eng.), Ph.D., M.I.E.E., who has been a member of the academic staff of the Faculty of Engineering, University College, London, since 1925 and is now Pender Professor of Electrical Engineering at the college, has been elected a Fellow of the American I.R.E. The citation reads: "For contributions to engineering education, telecommunication, and high-frequency techniques." At the beginning of the war Professor Barlow, who is 57, was at T.R.E. and subsequently became superintendent of the radio department at R.A.E., Farnborough. He has served on a number of Government boards and councils including the Radio Research Board, to which he was appointed in 1948.

For his contributions to the development of air traffic control systems, **J. Fenwick,** a senior signals officer in the Ministry of Transport and Civil Aviation, has been awarded the British Silver Medal of the Royal Aeronautical Society. He is the senior telecommunications officer in charge of the Southern Air Traffic Control Centre adjacent to London Airport and has been responsible for the design of the radar simulator which is used by the Ministry for the rapid training of control officers in radar control. Mr. Fenwick has also designed and engineered the whole of the radar display systems for the new Southern Air Traffic Control Centre.

C. P. Fogg has been appointed head of the ground radar department at the Radar Research Establishment of the Ministry of Supply, at Malvern, where, for the past five years, he has been superintendent of basic techniques. He joined the staff at the Bawdsey Research Station in 1937 and in 1939 was made leader of a group responsible for research and development of radio receivers. He went to Malvern in 1945 and was at one time in charge of the three divisions working on transmitters, aerial test gear and receivers and display. He is 42.

W. M. York, who has been in charge of Ekco publicity for the past twenty-four years and has been an executive director of the company since 1951, has been appointed commercial director with a seat on the Board. He recently visited the Ekco organization in India.

M. M. Macqueen, manager of the radio and television department of the G.E.C. since 1930, has accepted the chairmanship of the British Radio Equipment Manufacturers' Association for the third successive year at the express wish of the members of the council. He was also chairman in 1949 and 1950. Mr. Macqueen, who joined the G.E.C. in 1923, has represented the industry on a number of committees including the Radio Rearmament Advisory Committee set up by the Government in 1951 for liaison between the Ministry of Supply and the radio industry.

M. M. MACQUEEN

R. T. LAKIN

H. S. Jewitt has left Decca Radar, where he was senior engineer in charge of the receiver design group at the research laboratory, Tolworth, Surrey, and emigrated to the U.S.A. to join the Raytheon Manufacturing Company, of Boston, Mass. Before joining Decca in 1952 he was working on pulse circuitry at Ferranti's after graduating as B.Sc.(Eng.) from Queen Mary College, London University, in 1949. It will be recalled that Mr. Jewitt won one of the R.I.C. technical writing premiums for his articles on i.f. amplifiers published in *Wireless World* in 1954. He is 33.

R. T. Lakin, A.M.I.E.E., A.M.Brit.I.R.E., chief research technologist with the Whiteley Electrical Radio Company, and **H. W. Read,** London manager, have been appointed to the board of directors. Both of them have been with the company for more than twenty years. In 1953 Mr. Lakin was appointed M.B.E. for his scientific contributions during the war.

K. M. McKee, B.Sc., A.M.I.E.E., formerly a senior engineer for five years at the E.M.I. Research Laboratories, has joined the Geo. Tucker Eyelet Company, where he is in charge of the technical and commercial development of a range of automatic component assembly machines for use in printed circuitry.

L. S. King, contributor of the article in this issue on the Band I/III crossover network, has been with Standard Telephones and Cables for many years and for over thirty years has been part-time lecturer at the West Ham Municipal College and the Northampton Polytechnic, London, E.C.1.

A. H. Hooper, who contributes an article in this issue on the graphical derivation of radio refractive index, is a meteorologist. Among his other interests is experimental amateur radio and recent private investigations have included the feasibility of routine assessments of conditions for v.h.f. propagation through the troposphere.

OBITUARY

Stanley Whitehead, M.A., D.Sc., M.I.E.E., F.Inst.P., director of the British Electrical and Allied Industries Research Association (E.R.A.) for the past ten years, died on May 5th, aged 54. Dr. Whitehead, who had been a member of E.R.A. since 1925, was for some years chairman of the International Special Committee on Radio Interference (C.I.S.P.R.) and was a member of the advisory committees appointed by the P.M.G. to investigate radio interference from ignition systems and small motors. Since 1954 he had been joint honorary secretary of the Parliamentary and Scientific Committee.

Maurice G. Hammett, M.I.E.E., chief engineer of the electronics division of Murphy Radio since September, 1953, died early in May at the age of 48. For some time before joining Murphy's he was with the Ekco organization as chief engineer of the electronics division at Malmesbury. He started his engineering career with E.M.I.

WHAT THEY SAY

"Tall oaks . . .".—"Looking round this room I am struck by the thought that not much more than fifty years ago the entire manpower engaged in the radio industry numbered fewer than those representatives of it present here to-day. Indeed, it would be true to say that when my father [Guglielmo Marconi] first came to England he *was* the radio industry of the time!"—Marchese Giulio Marconi at the Radio Industries Club's 25th anniversary luncheon.

Incognito.—"Hitherto it has been the practice to classify this new electronics industry as an off-shoot of the long-established electrical industry, and as such it possesses no separate identity and no official classification. Statistically the electronics industry does not exist despite the fact that it employs some 250,000 people with an annual turnover exceeding £250 million and exports valued at over £60 million."—Sir Robert Renwick (president) at the 10th anniversary dinner of the Radar Association.

Obsolescent L. & M.W.?—"The B.B.C. is rapidly carrying through a complete conversion from long and medium waves to v.h.f."—Sir Ian Jacob, addressing the C.C.I.R. delegation.

IN BRIEF

Receiving Licences.—The overall increase in sound and television licences in the United Kingdom during the year ended April 30th was 278,533 bringing the total to 14,295,980. During the year television licences increased by 1,231,453 to 5,812,178 and sound licences decreased by 952,920 to 8,483,802.

The B.B.C. has selected the site for its **Cumberland television station** which it is planned to bring into operation by the end of next year. It will be built at Sandale which is 1,200 feet above sea level and some 14 miles south-west of Carlisle. The station will also be used for v.h.f. sound broadcasting when the system is extended to that part of the country.

B.B.C.-I.T.A. Co-siting.—Replying to a general question in the House of Commons on the sharing of masts by the B.B.C. and I.T.A., the Postmaster General announced that this will definitely be done in Cumberland.

I.T.A. Test Transmissions.—The Belling-Lee mobile television transmitter, which has done yeoman service in radiating test transmissions from London, Lichfield and Winter Hill during the building of the I.T.A. stations, has now been acquired by the Authority. In July it will be on Emley Moor, the site for the Yorkshire transmitter, for test transmissions with an e.r.p. of 1 kW.

R.I. Club Jubilee.—The 25th annual report of the Radio Industries Club (London) records a record membership of 885. At the anniversary luncheon on May 29th Sir Harold Bishop (B.B.C. director of engineering) handed over the presidency of the Club to Eric K. Cole who introduced the trade name Ekco in 1922 when he produced—at the rate of five or six a week—a two-valve battery receiver.

B.R.E.M.A. Council.—Member-firms whose representatives will constitute the council of the British Radio Equipment Manufacturers' Association for the ensuing year are: Balcombe (E. K. Balcombe); Bush (G. Darnley Smith); Cole (G. W. Godfrey); Cossor (J. S. Clark); Ferguson (S. T. Holmes); Ferranti (E. Grundy); G.E.C. (M. M. Macqueen, chairman); Gramophone Co. (F. W. Perks, vice-chairman); Kolster-Brandes (P. H. Spagnoletti); Philips (A. L. Sutherland); Pilot (H. L. Levy) and Ultra (E. E. Rosen).

The annual contests for **radio controlled models,** organized by the International Radio Controlled Models Society, will be held in the Midlands on August 5th, 6th and 7th. The contest for aircraft will be held on the 5th at the aerodrome, Wellesbourne Mountford, near Stratford-on-Avon, Warwickshire, and that for boats at Bournville, Birmingham, on the two following days. Further particulars and entry forms are obtainable from H. Croucher, 27 St. John's Road, Sparkhill, Birmingham, 11.

J. F. DOUST
(See page 253)

N.W. Germany.—With the reorganization of the broadcasting service in what was the British zone of occupation in Germany, the Nordwestdeutscher Rundfunk has been disbanded. There are now two sound broadcasting organizations in the zone: the Norddeutscher Rundfunk (N.D.R.) with its headquarters in Hamburg and the Westdeutscher Rundfunk (W.D.R.) centred on Cologne. There is, however, a co-ordinated television service for the zone, in which there are twelve television stations, operated by the Nord- und Westdeutscher Rundfunkverband (N.W.R.V.).

The Radio Industry Council is planning to hold a **Scottish Radio and Television Exhibition** at the Kelvin Hall, Glasgow, in mid-May next year. Although since the war there have been two Glasgow radio exhibitions organized by the Scottish Radio Retailers' Association, this will be the first manufacturers' show in the city since 1935.

Northern Electronics Show.—The 11th annual electronics exhibition, organized by the Northern Division of the Institution of Electronics, will be held at the College of Technology, Manchester, from July 12th to 18th (excluding Sunday 15th). It will open on the first day at 2 p.m. and on subsequent days at 10 a.m. and will close daily at 10 p.m. except on Saturday when it will be 6 p.m. During the course of the exhibition, which will include research and manufacturing sections, some fifty lectures will be given and sixteen films shown. Admission tickets will be obtainable free from exhibitors or from W. Birtwistle, 78 Shaw Road, Thornham, Rochdale, Lancs., from whom lecture programmes are obtainable. Separate tickets are being issued for each lecture.

No 1956 Amateur Show.—It has been decided by the Radio Society of Great Britain not to hold an amateur exhibition in London this year. The attendance at last year's exhibition was down by nearly 20 per cent on that of the previous year and the Society incurred a loss of about £70. The proposal that this year's exhibition should be held in the provinces—Manchester, Birmingham or Bristol—has not materialized.

E.M.I. College of Electronics has been adopted as the title of the department of E.M.I. Institutes which provides full-time day courses in radio and electronic engineering. Over 200 students are now attending the full-time courses at the college which offers eighteen scholarships for science sixth-formers to undertake the four-year course in electronic engineering and two scholarships for the three-year telecommunications course for which the entrance standard is G.C.E. (ordinary level). Full particulars of the scholarships and the courses, which begin in the autumn, are obtainable from the College, 10 Pembridge Square, London W.2.

A summer school on communication theory, modulation and noise is being held at Birmingham University from July 1st to 13th. It is intended for engineers in the electrical communications and allied fields who have taken their degrees without much theoretical work in communications and for those who feel the need for a "refresher" course. The fee is £8, excluding accommodation. Particulars are obtainable from the Director of Extra-Mural Studies, The University, Edmund Street, Birmingham, 3.

BUSINESS NOTES

G.E.C.-B.T.M. Collaboration.—The General Electric Company and the British Tabulating Machine Company are to collaborate in the design of "competitively priced computers and like equipment to cover the field in which such devices can be employed, and particularly the sphere of office machinery and automation." They are jointly to form a new company for this purpose.

RCA Photophone, Limited, which was established in this country in 1929 by the Radio Corporation of America primarily for the introduction of the Photophone system of sound recording for pictures, has changed its name to **RCA Great Britain, Limited.** The company is now the sole U.K. distributor for all products of the parent organization. Incidentally, the fiftieth anniversary of the formation of the Radio Corporation of America was celebrated at the end of February.

Ekco car radio, Model CR152/K, has been approved by the Rootes Group as standard or alternative optional equipment for their new Hillman "Minx" special saloon, de luxe saloon and convertible cars.

An industrial television camera has been installed by **Pye** as part of the new wind-tunnel of the Aircraft Research Association at Bedford. Direct observation through the usual plate-glass window is not possible in this tunnel because of the special method of lining the walls.

Included in the **Marconi** equipment being installed in the recently built *Manchester Venture* and *Manchester Vanguard,* which will trade between Manchester and the Great Lakes, is a new ship-to-shore radio-telephone set. Known as the "Seaway," it was designed by the Canadian Marconi Company to meet the requirements of the Canada/U.S. Great Lakes Treaty.

The new factory at Uxbridge opened by **Alfred Imhof, Limited,** instrument case makers, provides over 25,000 square feet of production space. The other factories at Islington and Thornton Heath will remain in use but the drawing offices and development section are now accommodated at Uxbridge.

This year marks the coming-of-age of **Painton and Company,** of Kingsthorpe, Northampton, manufacturers of a wide variety of components and accessories, including attenuators and faders.

The distribution of **Goodmans** loudspeakers and acoustical resistance units in Scotland is now undertaken solely by Land, Speight and Company, of 2 Fitzroy Place, Sauchiehall Street, Glasgow, C.3.

A by-product of the Redifon GR174 a.m. f.m. radio-telephone equipment is that it gives the marine user the opportunity to overcome interference should other vessels nearby be operating on a.m. on the same frequency. The combined a.m./f.m. 20-channel radio-telephone is shown installed in a deep-sea trawler.

Amplivox celebrates 21 years of hearing-aid manufacture this year. The Company, of which A. Edwin Stevens is governing director, also manufactures microphones and headphones.

In the refurbished demonstration centre recently opened by **Philips**, at Century House, Shaftesbury Avenue, London, W.C.2, a "hi-fi" demonstration room is provided in which elaborate switching arrangements permit various combinations of loudspeakers and amplifiers to be compared.

Birmingham Sound Reproducers are extending their factory space in Londonderry, Northern Ireland, to give a total of 180,000 sq. ft. B.S.R., who make the Monarch record changer and other gramophone units, have their head office and another factory at Old Hill, Staffs.

A.B. Metal Products are extending their Glamorgan factory to give an additional 50,000 square feet of manufacturing space.

Savage Transformers Limited, makers of the well-known "Massicore" transformers, have added an office block to their factory at Devizes, Wilts. This has released factory space for production.

Gate Electronics, Limited, have moved into their new factory in Tudor Grove, London, E.9. (Tel.: Amherst 8484.)

Electrical Instrument Repair Service, of 329, Kilburn Lane, London, W.9, have changed their name to E.I.R. Instruments Limited.

New offices and showrooms in Lancashire House, 9 South Street, Manchester 2, have been opened by **Marconi Instruments.**

OVERSEAS TRADE

Poland.—On their stand at the International Fair in Poznan (June 17th to July 1st) Kelvin and Hughes will be showing echo sounders, industrial equipment, miniature motors and also sound recording and reproducing equipment manufactured by their associates, Simplex-Ampro, Limited.

Export enquiries for the Sinfonia and President tape recorders and amplifying equipment manufactured by Phillips and Bonson, Limited, should now be addressed to Barnett Shipping and Export Company, St. Magnus House, 25 Monument Street, London, E.C.3.

Italy.—An order for £20,000 worth of equipment, including decade oscillators, telegraph distortion measuring gear and transistor test equipment, has been received by Winston Electronics, Limited, from the Italian Government.

Australia.—A three-camera O.B. vehicle has been supplied by Pye to the Australian Broadcasting Commission.

U.S.A.—William B. Allen, of 1601 Orleans Ave., New Orleans, Louisiana, wishes to get in touch with a United Kingdom manufacturer making a low-priced transceiver operating on 465 Mc/s. It should operate on a 6- or 12-volt car battery and have a range of ten miles or more.

Yugoslavia.—Pye telecommunication and television equipment and scientific instruments manufactured by their associates, W. G. Pye and Unicam, are to be shown at the Scientific Exhibition at Ljubljana, from August 4th to 12th.

U.S.A.—Olson Radio Warehouse Co., 275 East Market Street, Akron, Ohio, are interested in importing radio components, accessories, valves, etc., from the United Kingdom.

Honduras.—A. Idiaquez y Cia., of Tegucigalpa, D.C., are interested in importing, to sell in their own stores in Tegucigalpa and San Pedro Sula, a small battery-operated domestic receiver, preferably in a wooden cabinet.

Unconventional F.M. Receiver

Designed for Local Station Reception Using Pulse-Counter Discriminator

By ***M. G. SCROGGIE****, B.Sc., M.I.E.E.*

THE design of this receiver from the i.f. output onwards (comprising limiter, discriminator and de-emphasis circuit) has been discussed in detail in the April issue. So far as the rest of the receiver is concerned, the most significant feature of this discriminator is that its working frequency has to be of the order of 150 kc/s. To meet this requirement there are two alternative methods. One is to use a conventional f.m. receiver as far as the i.f. output and then interpose a second frequency changer to reduce the standard 10.7 Mc/s i.f. to 0.15 Mc/s. This procedure would combine the usual standard of sensitivity with the low distortion and absence of alignment that are the chief attractions of the pulse-counter discriminator. It may be that in some circumstances they would be worth the extra complication of the second frequency changer.

The other alternative is to accept 150 kc/s (or thereabouts) as the i.f. This extends the no-alignment benefit to the i.f. amplifier also, and the receiver as a whole is simple, but its sensitivity is low. However, where an adequate field strength exists this is no disadvantage.

In the interval since the publication of the discriminator article some readers seem to have been troubled in their minds about the problem of amplifying a bandwidth of 150 kc/s centred on an i.f. of about the same figure. As it happened, however, no difficulty at all was experienced with the i.f. amplifier; nearly all the trouble occurred in connection with the frequency changer.

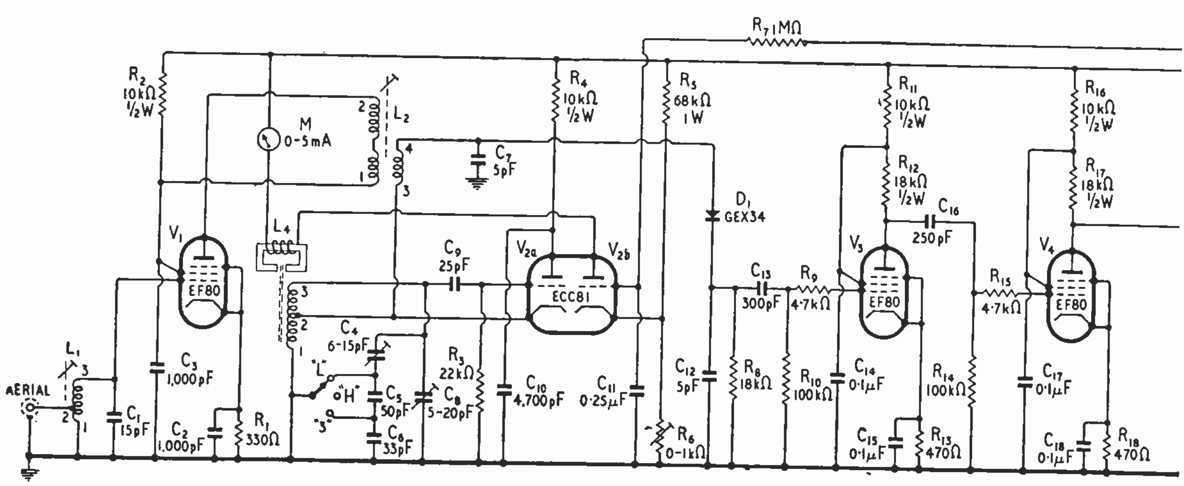
Fig. 1 shows the complete circuit diagram, from which it will be seen that two resistance-coupled i.f. stages are used in front of the limiter. The resistors in series with the grids provide, in conjunction with the valve input capacitances, attenuation of the upper frequencies, eliminating v.h.f. components received from the frequency changer and limiting the i.f. bandwidth at the top end, improving stability. There is of course further top cutting by the valve output capacitances shunting the coupling resistances. Fig. 2 is the amplification frequency characteristic. With reasonable care in the layout of the i.f. and discriminator stages and the wiring of the decoupling capacitors, it is unnecessary to use inter-stage screening.

The signal required at the grid of V_5 is at least 3V and preferably about 10V. The combined voltage gain of V_3 and V_4 is about 4,000, so the input to V_3 should be at least a millivolt or two.

The Frequency Changer

To reduce the incoming v.h.f. signal to 150 kc/s in one step requires the heterodyne frequency to differ from the signal frequency by only about 0.17%. This is small enough to suggest the use of a self-oscillating coupling circuit between r.f. and i.f. stages, but although the idea might be worth following up it was rejected on the ground of liability to radiate. Second-harmonic operation, though it would greatly reduce this liability, was also rejected, in the interests of televiewers on Channel 1, who are somewhat vulnerable to an oscillator fundamental frequency in the 44 to 47-Mc/s band! Third-harmonic operation puts the fundamental in a less awkward band and at the same time still further reduces radiation, so was adopted. An incidental advantage of the lower oscillator frequency, especially valuable if automatic frequency correction is

Fig. 1. Complete circuit diagram. The portion shown in Fig. 9 of the previous article appears on page 259. Coil L_3 is the tapped coil below L_4.



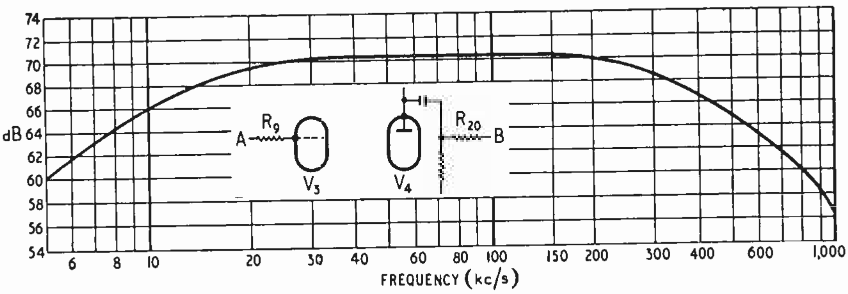


Fig. 2. Amplification/frequency characteristic of the i.f. amplifier, measured between A and B.

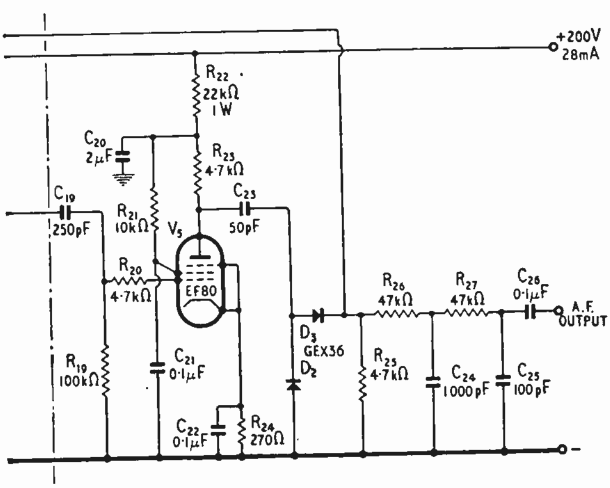
Frequency (Mc/s)	Station	Programme
88.1	N.H. Tor	
.3	S. Coldfield	
.5	Pontop Pike	
.7	Medlrum / Blaen Plwy	
.9	—	
89.1	Wrotham	
.3	Holme Moss	Light
.5	—	
.7	Norwich	
.9	Wenvoe	
90.1	Divis	
.3	N.H. Tor	
.5	S. Coldfield	
.7	Pontop Pike	
.9	Meldrum / Blaen Plwy	
91.1	—	
.3	Wrotham	Third
.5	Holme Moss	
.7	—	
.9	Norwich	
92.1	Wenvoe	
.3	Divis	
.5	N.H. Tor	
.7	S. Coldfield	
.9	Pontop Pike	
93.1	Meldrum / Blaen Plwy	Home
.3	—	
.5	Wrotham	
.7	Holme Moss	
.9	—	
94.1	Norwich	
.3	Wenvoe	
.5	Divis	

employed, is the greater latitude in the design of the oscillator. Against this, the efficiency of frequency changing is about one-third less than with second-harmonic and two-thirds less than with fundamental.

For domestic broadcast reception by all and sundry, station selection by a continuously variable tuning control belongs, in the writer's opinion, exclusively to the archaic era. What is required is an instantaneous switch-over from one correctly tuned programme to another. Especially is this true of f.m., to the tuning of which the non-technical listener has even less of a clue than usual, and where the provision of switch tuning is facilitated by the fact that all the programmes to be received come from a group of three transmitters having the same location, power and frequency spacing.

The pattern is shown in the accompanying Table, where the three transmitters comprising any group can be picked out by vertical movement of the three-pronged fork seen on the left, with a fixed inter-prong spacing of 2.2 Mc/s. (Incidentally, it is helpful to remember that the Light programme comes on the *Lowest* frequencies and the Home programme on the *Highest*.)

The tuning of the r.f. stage can be made broad enough to cover all three frequencies in any group and still give a useful gain, so for simplicity this was done, leaving the oscillator circuit as the only one to be varied. Substantially better sensitivity and selectivity can be obtained by variable r.f. tuning, if one cares to provide it. Whatever were done in this respect, however, would clearly not be enough to discriminate appreciably against the second channel, which with a 150 kc/s i.f. is only 0.3 Mc/s off-tune. This comes half way between the carrier frequencies of two other stations, but both of them give heterodyne frequencies well within the pass band of the i.f. amplifier. That these do not cause interference is due to a combination of two things: the low sensitivity of the receiver, and the "capture effect" in f.m. reception. Interference is not appreciable in f.m. receivers unless the interfering signal approaches the same order of magnitude as that being received; but this happens only in fringe areas, for which the sensitivity of this receiver is insufficient. The only signals of comparable strength will therefore be the other two from the local station, and they come beyond the i.f. pass band on both heterodyne channels. Several types of frequency changer were tried, including all those shown in designs published in *Wireless World* during the last few years, but much the most satisfactory was a type not shown in any—the ordinary diode with separate oscillator, as used for microwaves. Attempts to involve V_3 in any part of the frequency-changing process—and especially oscillation—gave very inferior results. Whatever frequency changer is adopted, care must be taken to ensure the absence of squegging, which at an ultrasonic frequency can easily be mistaken for instability of the i.f. amplifier.



The oscillator is fairly conventional, and supplies about 6 to 7V in series with the signal from the r.f. stage to the diode D_1. Best results were obtained with a thermionic diode (type EA50) and rather surprisingly there was no trouble with hum from the "live" cathode. Nevertheless, there is obviously a lot to be said for a germanium diode in this position, with the exception that the performance at this frequency of those types that can take the oscillator voltage appears to be somewhat uncertain. The samples tested gave 50% to 70% of the output of the EA50, but because the best of these provided ample level at the limiter it was adopted for its convenience. The d.c. component across R_8 is about 1V less than the r.m.s. oscillator input to D_1.

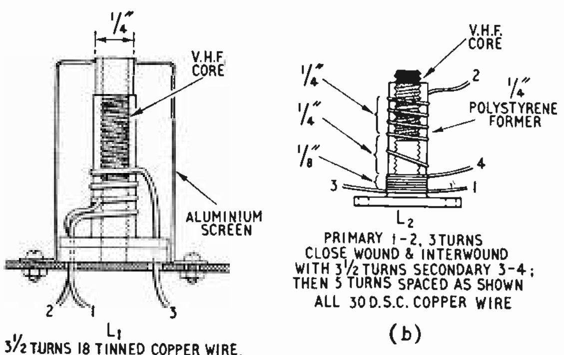
An important component in the signal chain is the coupling from the r.f. stage L_2. Of a large number of couplings tried, the one shown in Fig. 3 (b) is the most satisfactory. It is very similar to the one described by Amos and Johnstone in the October, 1952, issue. The rest of the r.f. stage, including the input coil L_1 which is also shown in Fig. 3, is quite conventional.

For the purpose of calculating the oscillator tuning circuit values for the three programmes, the difference between the signal frequency and three times the oscillator frequency can be neglected. Each of the outside signal frequencies (Home and Light) differs by 2.2 Mc/s from the middle one (Third), or 2.4%. This necessitates very nearly 5% capacitance difference. Theoretically the two steps are not exactly equal, but the difference is small enough to neglect. The most obvious arrangement is that shown in Fig. 4(a). A suitable value for the total capacitance needed for one-third of any of the Third programme frequencies is 36 pF, for which the corresponding inductance is about 0.76 μH. Of this, the stray capacitance is likely to be about 12 pF, leaving 24 pF to be provided. Five per cent of 36 pF is 1.8 pF, so the capacitances required are approximately as shown. The switched capacitances are so small as to be complicated by the stray capacitances of the switch, and also (because the whole oscillator voltage is applied) by its losses.

Fig. 4(b) shows an alternative that avoids these disadvantages, for the switched capacitances C_5 and C_6 can be made relatively large and consequently low in potential. If the capacitance steps are denoted by δC, then

$$C_5 = \frac{C_4(C_4 - \delta C)}{\delta C}$$

$$C_6 = \frac{(C_4 - \delta C)(C_4 - 2\delta C)}{\delta C}$$



Above: Fig. 3. Constructional details of the v.h.f. tuning coils, L_1 and L_2.

Substituting 1.8 for δC, and trying $C_4 = 10$, we get $C_5 = 45.5$ and $C_6 = 29$. Now with a fixed L_3 the correct tuning for the Home programme can be pre-set by adjusting C_8. The precise values of δC for the other two programmes depend on the total tuning capacitance and hence on L_3, and as 0.76 μH is only a target value for L_3 there is need for some adjustment of δC to ensure correct tuning of the other programmes. One does not want to have to pre-set all three capacitors in the switching group; the most convenient is C_4, but does adjustment of it vary both steps equally? Let us assume for the moment that it does. Then the calculated values of C_5 and C_6, 45.5 pF and 29 pF, are rather odd for fixed components. But putting $C_5 = 50$ we get $C_4 = 10.4$ and $C_6 = 33$, which if not directly available can be made up from 100 and 50 in series. Assuming these values, and plotting the two capacitance steps against C_4 we get Fig. 5, which shows that although exact equality is not maintained the values of δC can be varied over sufficiently wide limits without serious discrepancy.

Automatic Frequency Connection

To correct for any such error, and to render unnecessary any particular care in the choice of components or other precautions for counteracting oscillator drift, automatic frequency correction was adopted. There are various possible ways of providing it, such as the milliammeter-operated system described by C. H. Banks in the February issue. For the fun of it, the writer experimented with the core-saturation system shown. The oscillator coil L_3 is wound on a small ferrite core placed in the gap of an electromagnet L_4, which is energized by the anode current of the spare half of V_2. A z.f. controlling voltage exactly proportional to the i.f. is available at the output of the discriminator; as shown in the previous article, the voltage corresponding to 150 kc/s is 2¼. R_6 is adjusted so that with an equal voltage applied from an independent source to the grid of V_{2b} the anode current is a convenient value (in this case about 2½ mA, mid-scale on the tuning indicator M) and the oscillator trimmer is adjusted so that with the switch at "H" the Home programme is tuned in to give 2¼ V output from the discriminator. If this cannot be done with a total capacitance reasonably near the design value, the number of turns on L_3 must be altered. When satisfactory, the discriminator output can be substituted for the independent source. The i.f. and a.f. must of course be filtered out, for which purpose R_7 and C_{11} are provided.

To hold the tuning, an increase in i.f. must affect the inductance L_3 in such a way as to reduce the

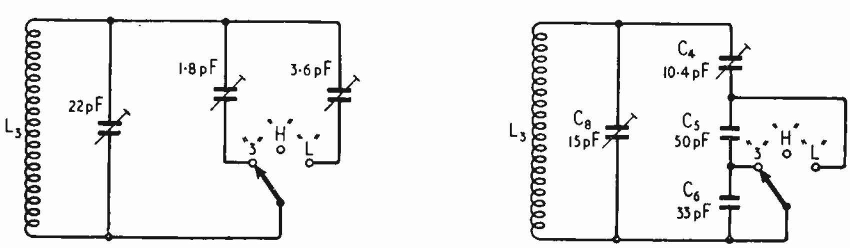


Fig. 4. Two alternative arrangements of the programme selector switch, of which (b) is preferred. The capacitance values shown are calculated for $L_3 = 0.76$ μH and stray capacitance 12 pF.

i.f., and a little consideration will show that for this to happen it is necessary for the oscillator to be tuned so that its third harmonic is 150 kc/s *lower* in frequency than the signal.

The oscillator inductor L_3 consists of 12 turns of 30 d.c.c. wire wound more or less toroidally on a core of type B4 Ferroxcube, list No. FX.1595, which is ¼in square with a central hole and resembles a ⅛-in square nut. For experimental purposes a 10,000 Ω Type 3,000 P.O. relay was used as the polarizing system L_4, with the armature and springs removed and the circular pole piece filed on one side to make a ¼-in parallel gap to take the L_3 core. This gave an extremely tight control of tuning with only about 0.4 mA normal polarizing current, but the low reluctance of the closed magnetic circuit was found to be very vulnerable to stray magnetic fields frequency-modulating the oscillator at 50 c/s and so causing hum. This could presumably be eliminated by placing the whole relay in a Mumetal screen. Another disadvantage was that the ferrite is subject to hysteresis, and if the polarizing current accidentally went outside its normal control limits (say in the setting-up) it was necessary to switch off and start afresh, in order to ensure that the core was working on the rising-from-zero part of its magnetic cycle.

Because of its lower hysteresis (and higher permeability) a ring core of B2 Ferroxcube was tried as an alternative to the B4, but its higher losses stopped oscillation at 30 Mc/s. (Incidentally, even B4 would be unworkable at the oscillation frequency of a fundamental frequency changer.)

By turning the oscillator core at right angles as shown in Fig. 6 and wedging it in position with small pieces of polystyrene as magnetic gaps, the liability to hum was greatly reduced, and the looser a.f.c. and larger polarizing current (2½ mA) reduced the extent to which the tuning could be shifted by hysteresis. It is still necessary to keep mains transformers a foot or two away and check their orientation, and of course the anode supply to V_{2b}—as in fact to the other valves—must be thoroughly smoothed. The actual h.t. voltage is not at all critical and can be varied over wide limits.

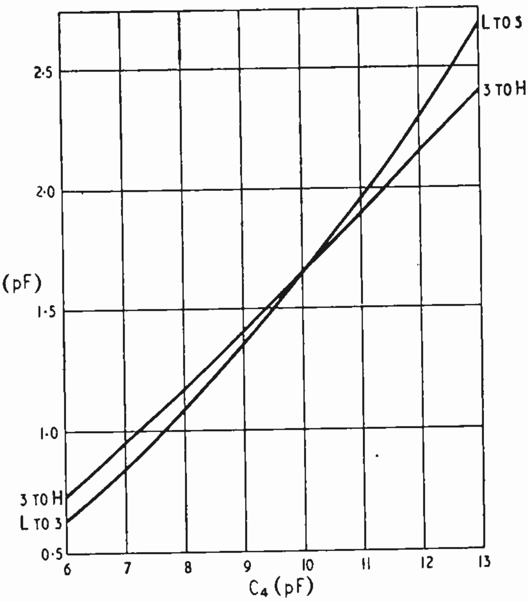


Fig. 5. With the values of C_5 and C_6 as in Fig. 4(b), the jumps of capacitance between the positions of the programme switch vary with C_4 as shown.

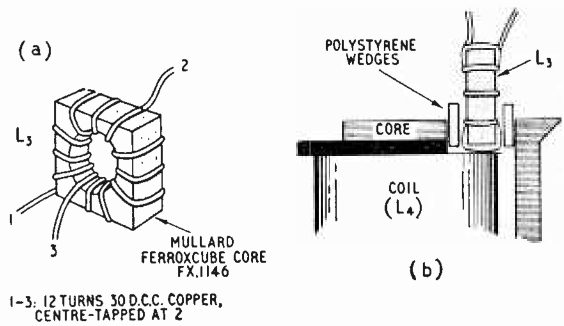


Fig. 6. Details of the experimental oscillator tuning coil and polarizing magnet for the automatic frequency correction system described.

Adjusting the Oscillator

The a.f.c. system and the oscillator circuit having been set up to receive the Home programme as described, C_4 must be set to give correct tuning of the other two programmes on working the switch. Correctness is indicated by the reading of M remaining constant, or nearly so. If switching over to Light makes the reading rise, C_4 is too large; with the switch back at Home it should be slightly reduced and at the same time C_8 increased to keep the total as before. The process is repeated until tuning is reasonably correct for both programmes. If the values of C_5 and C_6 are close to the specified values, the tuning of the Third should then be good enough. Because with the pulse-counter discriminator the i.f. is not at all critical, it is not a serious matter if the control current changes even half a milliamp between stations, but it does reduce the margin of a.f.c. available for dealing with warming-up drift, etc. The meter M is not essential as a permanent feature of the receiver, but is useful for indicating departure from correct adjustment. This programme switching system works extremely well, giving an instant change-over with little or no click, and of course the limiter ensures that all three programmes give exactly the same a.f. voltage for a given depth of modulation. Incidentally, the output is sufficient to drive a Leak TL/10 amplifier fully without pre-amplifier.

For adjusting the cores of L_1 and L_2 and checking the adequacy of the i.f. signal, a valve voltmeter from grid of V_5 to earth is very helpful; if a proper instrument is not available, one can be extemporized, or an oscilloscope used. The low intermediate frequency facilitates such measurement. To tune L_1 and L_2, the cores should be adjusted to peak on the Third programme; then slightly staggered (one up and one down) to give approximately equal readings on Home and Light.

The receiver here described is not to be regarded as a final design but rather as a suggestion for interesting experiment. The relay as a polarizing magnet is only a makeshift that happened to be handy; no doubt a much smaller component could be designed for the purpose. Or one might prefer an altogether different system of a.f.c. Then the question of how this type of receiver compares with the conventional as regards freedom from noise and interference has not been fully investigated. The writer suspects that on a level comparison it might be somewhat less good, but, just as in the matter of its inferior selectivity, it can afford to give something away because of its lower sensitivity. The hybrid type—the double superheterodyne—combines the advantages of both. Lastly, remembering that one of the chief claims of the pulse-counter is its low distortion, one must consider possibilities of distortion elsewhere. One such possibility is inequality of amplification over the i.f. band. Provided that the amplitude at the input to the limiter does not vary beyond the limits of the almost perfectly level part of the output voltage characteristic shown in Fig. 8 of the previous article, there should be no appreciable amplitude modulation to affect the discriminator; but any amplitude variations at the input might slightly vary the ratio of positive to negative "half" cycles, and in that way introduce a certain amount of phase modulation, which would distort the f.m. waveform. This too has not been fully investigated, but a rough calculation seems to indicate that with the 1 dB drop at the upper peak of 100% modulation from 150 kc/s shown in Fig. 2 the distortion would be small. It could be reduced (along with the very slight discriminator distortion indicated in the previous article) by reducing the i.f. to, say, 100 kc/s; but it is necessary to leave enough margin for there to be no chance of the i.f. drifting so low as to clash with the a.f., for that causes most evident distortion.

LONG RANGE ON V.H.F.

DURING recent months the solar activity, which was expected to increase at a rapid rate, has done so even more rapidly than was anticipated. The response of the F_2 layer to this increase in the activity of its producing agent has been most marked. Its ionization has risen rapidly, and, during the period February-April, 1956, the monthly means of the noon F_2 critical frequencies as measured at Slough were of the order 9-10 Mc/s.

So far as radio propagation is concerned, therefore, it may be said that quasi-maximum conditions; i.e., those associated with a sunspot number greater than 100, already exist. And they are producing some interesting, if sometimes unwanted, results. The higher frequencies in the h.f. band are being regularly propagated by the F_2 layer over long distances, whilst those in the lower part of the v.h.f. band are similarly affected.

The prediction curves published monthly in *Wireless World* under "Short-Wave Conditions" have given some indication of the very high frequencies expected to become usable; for example, for 25% of the total time over the paths to Montreal and Johannesburg. In point of fact these have been somewhat exceeded. U.S.A. amateurs on 28 Mc/s were receivable in this country for almost half the days of March and April, whilst mobile radio signals on 30-35 Mc/s were frequently received from southern U.S.A. From several places in South Africa and Southern Rhodesia, including Johannesburg, reports were received of the frequent reception—almost daily during late March and early April—of the sound channel of the London television service on 41.5 Mc/s. And very often the video channel on 45 Mc/s was receivable after a fashion.

The solar activity is likely to continue to increase at a rapid rate, though not, perhaps, without considerable fluctuations. The seasonal effect in the Northern Hemisphere is, however, at present tending to produce lower values of daytime F_2 layer ionization, and this may cause some slight reduction in the daytime m.u.fs towards midsummer. After that, however, the seasonal effect will be operating in the same direction as the increasing solar activity, and the m.u.fs are likely to rise to even higher values than at present. By the autumn, therefore, it is probable that the F_2 layer will be capable of propagating phenomenally high frequencies.

Apart from the general effect of the solar activity in increasing the ionization of the upper atmosphere there is another effect which is of importance in radio communication; i.e., as the activity increases there is an increase in the frequency and severity of ionospheric disturbances.

Ionospheric disturbances are of two distinct kinds, and we may deal first with the short-lived disturbance—lasting perhaps for an hour—known as a "Dellinger" fadeout. These are produced by solar flares (usually associated with sunspots) and result in a complete fadeout of short-wave signals on certain frequencies which traverse the daylit hemisphere of the earth. During the first six months of 1955 only two such fadeouts were reported and during the second half of the year only five. But during the period January-April, 1956, no fewer than twenty such fadeouts were reported, some of which were of great intensity. "Dellinger" fadeouts are likely to be relatively frequent during the next few years.

But it is the other kind of ionospheric disturbance, known as an ionospheric storm, which constitutes the major disruptive phenomenon in short-wave communication, since its effects often last for several days. This again originates in the sun and is thought to be due to the emission of a corpuscular stream which, travelling at a velocity of the order of 1,000 miles per second, enters the earth's atmosphere and produces, among other terrestrial effects, the ionospheric storm. Since the corpuscular streams appear often to occur in association with sunspots—sometimes reaching the earth about 30 hours after a solar flare—their frequency and severity can be expected to increase with the increasing solar activity. Indeed, there has already been a marked increase in the frequency of these disturbances.

The next maximum of solar activity is expected to occur sometime during the first half of 1957, and to be one of outstanding intensity. The F_2 m.u.fs may therefore be expected to continue to increase towards that time, but the occurrence of ionospheric disturbances to become more frequent and, perhaps, more prolonged and severe. This state of affairs should continue during the course of the International Geophysical Year, which is a circumstance favouring the acquisition of new knowledge during that period of intensified scientific effort.

T. W. B.

Characteristics of Fixed Resistors

This article is based on certain sections of the author's book, "Radio and Electronic Components, Vol. 1: Fixed Resistors," recently published by Sir Isaac Pitman and Sons at 28s.

Sources of Noise, Resistance at High Frequency and Some Measurements on Carbon Composition Resistors

By G. W. A. DUMMER, M.B.E., M.I.E.E

ALL resistors generate noise due to thermal agitation of electrons. This is known as Johnson noise, after J. B. Johnson who discovered the effect in 1927. The random motion of free electrons in a resistor is in equilibrium with the thermal motion of the molecules. This random motion is superimposed on the electron drift current arising from the potential difference across the resistor. This produces a fluctuating voltage, or noise, and its r.m.s. value is only proportional to resistance and temperature. It has a uniform spectrum of frequency distribution and if different bandwidths are selected for a fixed value of resistance the relation between generated noise and bandwidth for a 10-kΩ resistor will be approximately as given in Table I.

For a given bandwidth the relation between Johnson noise and resistance value is approximately as shown in Fig. 1. In general the higher the resistance value, the greater the noise.

Carbon composition resistors also generate noise due to the passage of current through them. This causes random changes in the material of which the resistor is made. This noise, sometimes called "Bernamont" noise (after Bernamont's work in 1934), is also greatly dependent on bandwidth. Unlike Johnson noise, however, which is uniform at all frequencies, current noise decreases with increasing frequency from below about 10 c/s up to at least the kilocycle range. The generated noise (E) is related to bandwidth in the following way:

$$E^2 \propto \frac{f_2}{f_1}$$

where f_1 and f_2 are the lower and upper frequency limits of the noise-measuring equipment or amplifier, so that for a constant bandwidth the square of this noise voltage is inversely proportional to the frequency.

Current noise is particularly dependent on the physical construction and materials used in the manufacture of the resistor and can vary greatly between resistors of similar types and even between those of the same value and manufacture. In general the higher the direct voltage across the resistor the higher the noise.

Another important factor is the size of the resistor in relation to its wattage; with a given value under similar current conditions the noise in a small resistor will be greater than that in a large resistor Fig. 2 shows the relative noise voltage developed in three different sizes of carbon composition resistors using an a.f. amplifier with a bandwidth of about 10 kc/s.

Measurements have shown that for carbon com-

TABLE I

Bandwidth of Amplifier or Measuring Equipment	Johnson Noise Voltage (r.m.s.) in μV (approximately)
1 kc/s	0.6
10 kc/s	1.5
100 kc/s	6.0
1 Mc/s	20.0
10 Mc/s	60.0

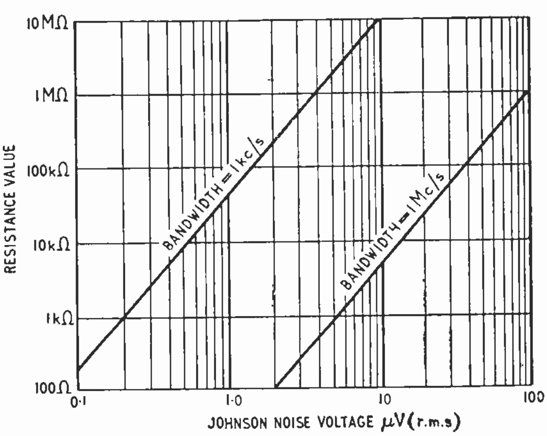


Fig. 1. Relation between Johnson noise and resistance value.

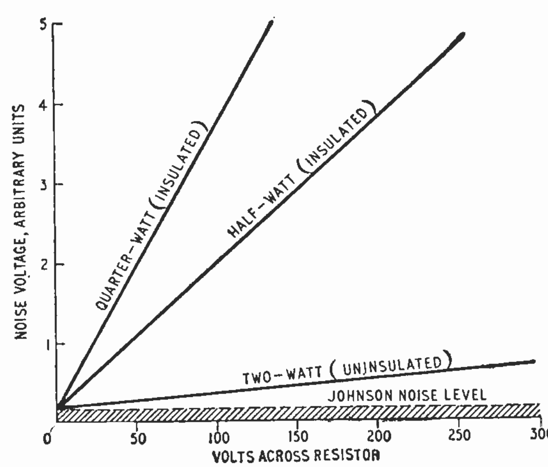


Fig. 2. Noise voltage from carbon composition resistors of the same value but different wattage ratings.

position resistors current noise increases linearly with current up to about 15 μA. With greater currents the noise curve approximates to a parabola. On a range of carbon resistors of different values up to 1.5 MΩ the current noise generated at normal voltages may vary approximately as in Table 2.

It is of interest to note that the total noise allowed in composition resistors for Service use is given by

$$2 + \log_{10}\left(\frac{R}{1{,}000}\right)\ \mu V/V$$

where R is the resistance in ohms. The noise-measuring equipment has a bandwidth of from 200 to 10,000 c/s. On this assumption the maximum amounts of noise allowed for various values of carbon composition resistors are given in Table 3. The third column in this table gives the maximum noise generated in cracked carbon resistors under equivalent conditions and is included here for comparison.

R.F. Performance.—A resistor must, by its construction, contain a certain amount of capacitance and inductance as well as resistance. A simple equivalent circuit of a resistor is that of Fig. 3. R is the "pure" resistance, C_1 may be considered the sum of many small distributed capacitances along its length and L is the lead, and any other inductance in the actual resistor. C_2 is the mounting capacitance when the resistor is included in a circuit.

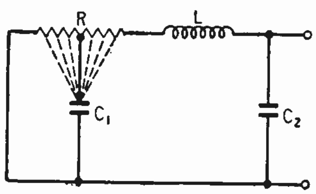


Fig. 3. Simple equivalent circuit of a resistor.

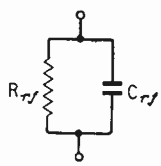


Fig. 4. Equivalent circuit of a resistor at a given frequency.

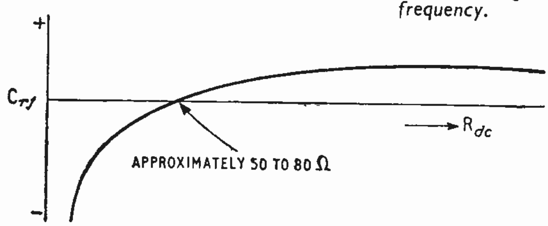


Fig. 5. Variation of C_{rf} with R_{dc}.

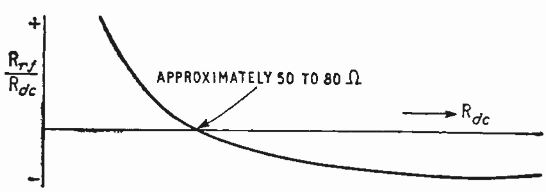


Fig. 6. Variation of R_{rf}/R_{dc} with R_{dc}.

At any one frequency the resistor can be represented by the simple parallel circuit of Fig. 4. Here C_{rf} may be positive or negative. For low values of resistance the variation of C_{rf} with resistance value takes the form of the curve in Fig. 5. The r.f. resistance, R_{rf}, again for low values of resistance, and plotted as the ratio of R_{rf}/R_{dc} against the value R_{dc} (d.c. resistance) is as shown in Fig. 6. It can be seen from this curve that, due to the series inductance, the r.f. resistance of very low value resistors rises above the d.c. values.

In general the r.f. resistance, or impedance, of all resistors decreases with an increase of frequency because of the distributed capacitance (C_1, Fig. 3) and the higher the value of the resistor the greater the fall in r.f. resistance. The performance of typical carbon composition resistors in values from 1 kΩ to 1 MΩ and at frequencies up to 40 Mc/s is shown in Fig. 7. Here the fall in resistance is plotted as the ratio of r.f. to d.c. resistance against increasing frequency.

Low-value resistors can be used as load resistors up to several thousands of megacycles and their behaviour at frequencies up to 1,500 Mc/s is shown in Fig. 8. A useful rough rule is that when Rf is less than 0.03 (where R is in MΩ and f in Mc/s) the r.f. resistance of a carbon composition resistor is within 10 per cent of the d.c. value.

Measurement of Noise.—The usual method of measuring the noise generated in a resistor is to feed the noise voltage into an amplifier of known gain and bandwidth and to detect the noise voltage on a thermo-couple voltmeter. The noise voltage is generated by applying a direct voltage across the resistor under test and a matching load in series.

Certain precautions must be taken to ensure that the extremely small voltage to be measured is not masked by stray pick up or instrument-generated noise voltages. Such precautions include screening of the input circuit, effective mains filtering and the use of low-noise valves in the early stages of the amplifier. Resilient valve mounting might be needed in the first amplifying stage to avoid microphony effects.

Measurement of R.F. Resistance.—There are two main systems of measuring the r.f. resistance of a resistor. One is of a more fundamental character, involving standards of capacitance (resonance

TABLE 2

Volts (D.C.) applied to resistor (V)	Approx. Current Noise (μV)		Johnson Noise (μV)
	Min.	Max.	
50	250	1,500	30–35
100	350	2,300	30–35
150	375	2,900	30–35
200	400	3,400	30–35
250	450	3,600	30–35
300	450	3,800	30–35

TABLE 3

Resistance Value	Maximum Noise (μV/V)	
	Carbon Composition Resistors	Cracked Carbon Resistors
1 kΩ	2.0	0.03
10 kΩ	3.0	0.15
100 kΩ	4.0	0.35
1 MΩ	5.0	0.50
10 MΩ	6.0	0.52

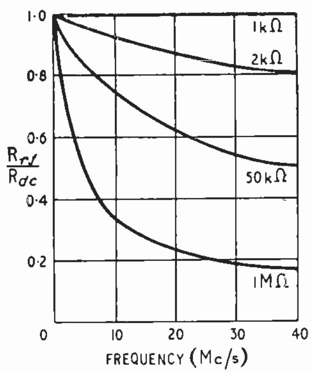


Fig. 7. Variation of R_{rf}/R_{dc} with frequency for different values of carbon composition resistors.

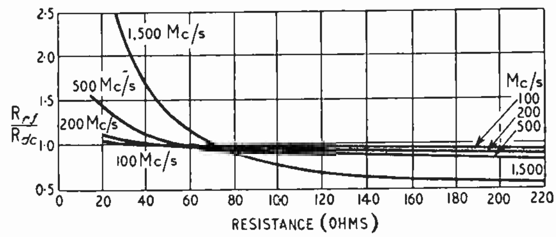


Fig. 8. Behaviour of low value carbon composition resistors at high radio frequencies.

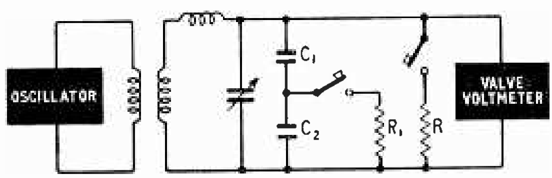


Fig. 9. Substitution method for r.f. measurements on resistors.

methods) or lengths of standard line (transmission-line methods). The other uses resistors as standards (bridge substitution methods). Resonance and transmission-line methods make no preliminary assumptions about the resistors to be measured; bridge methods are in most cases comparison systems. A simple substitution method can take the form shown in Fig. 9. R_1 is a low value "standard" resistor and R is the resistor to be measured. R and R_1 are brought alternately into the circuit until a value of R_1 is found which results in the voltage across the resonant circuit being the same with both R_1 and R. The unknown resistor (R) is then given by

$$R = R_1 \left(1 + \frac{C_2}{C_1}\right)^2$$

Transmission-line methods use slotted-line impedance-measuring equipment in which the terminating impedance of the line is determined by plotting the voltage standing wave pattern along the line. The length of slotted-line required is equal to half a wavelength at the frequency being used. The standard, with which the resistor whose r.f. resistance is required to be found is compared, is a length (or rather number of lengths) of accurately made short-circuited line. These "standard lines" are used to calibrate the measuring line. All measurements are then made in terms of lengths and the method is applicable up to about 3,000 Mc/s.

In all measurements on resistors at radio frequencies the method of mounting the resistor is important. The end-to-end capacitance of the resistor and the capacitance of the two leads to the resistor body are included in the total capacitance of the resistor being measured and the resistor should therefore be mounted as nearly as possible as it is when in use.

Summarizing, for a resistor to be suitable for operation at radio frequencies it should meet the following requirements:—

(1) Its dimensions should be as small as possible.
(2) It should be low in value.
(3) It should be of the film type.
(4) A long thin resistor has a better frequency characteristic than a short fat one.
(5) All connections to the resistor should be made as short as possible.
(6) There should be no sudden geometrical discontinuity along its length.

Origins of Radio Telephony

IN the April issue "Free Grid" enquired when radio telephony was first used. A Danish correspondent, G. Bramslev, happens to have been conducting some researches into this obscure branch of radio history, and sends some notes on his findings.

In Fessenden's U.S.A. Patent, No. 706,747 of September 27th, 1901, relating to generation of r.f. by means of an alternator, there was a claim covering the use of the apparatus for telephony, with a telephone receiver at the receiving station. However, no details of the receiver were given.

In 1903 Valdemar Poulsen of Denmark developed the oscillating arc, and in his Danish patent, No. 8208, of September 27th, 1904, stated that wireless telephony could be realized by putting a microphone in the oscillatory circuit and so modulating the amplitude of the oscillations.

Thus far, there were no claims to practical realization, and the first descriptions of actual experiments which Mr. Bramslev has been able to trace were published in 1906. In the German periodical *Elektrotechnische Zeitschrift* of November 15th of that year, Ernst Ruhmer wrote about radio-telephony tests he had conducted with a Poulsen generator, carbon microphone and an electrolytic detector in the receiver. The distance covered was only 30 yards. An interesting fact is that Ruhmer mentioned both amplitude and frequency modulation, and, apparently, fully understood the nature of both systems.

Fessenden, after having successfully demonstrated the use of his rotary r.f. generator, turned to practical experiments in telephony, and in December, 1906, transmitted both speech and music over quite long distances from his station in Massachusetts, U.S.A.

What Mr. Bramslev says about Fessenden is borne out by Blake's "History of Radio Telegraphy and Telephony" (1926), which states categorically "He was the first to use a microphone in the aerial circuit." Blake adds that, by the use of telephone relays of his own design, he demonstrated the possibility of connecting a land-line telephone to a radio-telephone station. According to a quoted statement of Fessenden's, speech was "transmitted over land-line to the station at Brant Rock and re-transmitted wirelessly by a telephone relay, received wirelessly at Plymouth, and there relayed on another land-line." In 1907 Fessenden successfully transmitted speech over a distance of 200 miles.

Undoubtedly both the r.f. alternator and the arc played important roles long before the valve was used.

MORE ABOUT NOISE

Answers to Some Questions

By "CATHODE RAY"

BEFORE going on with this subject we had better recapitulate last month's findings

" Noise " in the title is not a class of sound, but " unwanted energy " in a communication system. True, it can cause unwanted sound, but it can also cause unwanted marks on paper or unwanted flecks on a cathode-ray tube. Our particular inquiry is confined to what is called thermal noise—the energy of movement of free electrons in solid matter. This mechanical energy is what happens to the heat energy that has been put into the matter to raise its temperature from absolute zero (-273°C). And because (according to the first law of thermodynamics) there is a fixed rate of exchange between the two kinds of energy, and temperature is proportional to heat energy, noise is proportional to absolute temperature. And because (according to the second law of thermodynamics) one resistor connected in parallel with another cannot raise the other to a higher temperature than itself, it can easily be shown that the square of the noise e.m.f. in any part of a circuit must be proportional to the resistance of that part, and the noise current squared must be inversely proportional to the resistance. The same conclusions follow via a more complicated chain of reasoning (which I only outlined) from basic assumptions about electrons in matter, such as that in a uniform piece of material the free electrons are uniformly distributed, and that the conductivity of the material is proportional to the number of these electrons per unit volume. Because the number of electrons is enormous and their movements are completely random, the average net number moving in any particular direction can be calculated with considerable precision on a basis of probabilities, in exactly the same way as the average departure from exact equality of heads and tails when vast numbers of coins are tossed. The maximum noise power that any part of a circuit can deliver to another —which happens when, by another basic principle, the resistances of the two parts are equal—is the same for all resistances. It is kTB, where $k = 1.38 \times 10^{-23}$ (Boltzmann's constant), T is the absolute temperature (centigrade degrees above -273), and B is the frequency band in cycles per second. This is the key formula for thermal noise, and is due to Nyquist of negative feedback fame. Because the terminal voltage when a generator is supplying a matched load is half its internal e.m.f., it follows that the internal r.m.s. noise e.m.f. in any resistance R is $\sqrt{(4kTBR)}$.

Even if you have read the previous instalment and possess at least a glimmering of an idea why the noise voltage squared should be proportional to the absolute temperature and the resistance, and are prepared to accept the need for some constant to make the units right, you may be inclined to query B. All I said about it was that since the electron movements are completely random their fluctuations responsible for noise occur equally at all frequencies, so noise power is uniformly distributed over all frequencies. If you are not satisfied with that explanation—and I don't blame you if you aren't—you will have to refer to the books, because frankly I have yet to find any way of showing it that is both simple and convincing. Certainly the noise is greater the greater the frequency band accepted by the measuring device—that is what one would expect, and it is confirmed by experiment—but if the relationship is as per Nyquist it looks as if over an unrestricted frequency band the noise power is infinitely large, which no one will believe. I did point out that in practice the frequency band is never unrestricted, because any resistance has some stray capacitance which increasingly shunts it as the frequency goes up. But quite apart from that the assumptions on which Nyquist's formula is based cease to apply somewhere in the region of 6,000,000 Mc/s. So even if there were no other restriction of bandwidth the maximum noise power available from anything would be limited (at 27°C) to $1.38 \times 10^{-23} \times 300 \times 6 \times 10^{12} = 2.5 \times 10^{-8}$ watts $= 0.025$ microwatt, which is not going to blow anyone's head off.

Maximum Noise Power

The next question that may be forming in the mind is: why this emphasis on *maximum* available noise power? Shouldn't we be more interested in obtaining the minimum? True, but the minimum is obtainable by short-circuiting the resistor responsible for the undesired noise and thereby bringing it to zero. This would also bring the desired signals to zero so would be of no practical benefit. The condition for maximum noise, on the other hand, is normally the condition for maximum signals; and in any case it is a standard condition for ready comparison and it has the simplest formula—kTB.

Next, if the noise e.m.f. is proportional to the square root of the resistance, it would be greatest in material of the highest resistivity, which has least free electrons. A perfect insulator, with no free electrons, would presumably generate infinite noise voltage! Does this not contradict the theory that noise energy is the energy of free electrons? If one rashly pursues matters to infinity or zero, however, one must be prepared for absurdities. Any finite amount of power, however small, developed in a resistance, would result in an infinitely high voltage if the resistance became infinite. At the same time the current would become infinitely small. But there is no possibility that this extreme state of affairs could ever be reached in practice, because even the best insulators contain some free electrons; and moreover

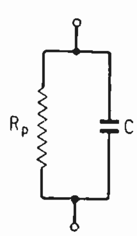


Fig. 1. This can be regarded as the simplest noise-making circuit, because any resistance is shunted by stray capacitance, such as the grid capacitance of the amplifier that reveals the noise.

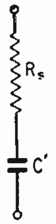


Fig. 2. This circuit is equivalent to Fig. 1 at any particular frequency, if the values are correctly chosen for that frequency.

the higher the resistance the narrower the frequency band, owing to the influence of stray capacitance.

And that is the cue for dealing with another possible cause of doubt: the effect on noise of circuit reactance. But before going on to that there are still some possible queries about resistance. Although I talked a lot last month about "resistors" as the source of thermal noise, it was not meant to imply that noise of that kind originated only in components sold in shops as resistors. Just as much noise arises in a 500-ohm transformer winding as in a 500-ohm resistor, except in so far as the resistance may change with frequency. That proviso raises the question of how, for purposes of noise calculation by means of Nyquist's formula, one reckons the resistance. Is it the resistance of the wire or whatever the circuit is made of, or does it include the effects of losses in iron cores, capacitor dielectrics, etc?

The answer is that in reckoning the part of the total noise coming within any narrow frequency band the resistance that counts is the value actually measured at that frequency. It would include the effects of all incidental losses—even radiation resistance. This may seem rather surprising, and it can lead to some quite involved arguments if one tries to explain theoretically how some of the more "fictitious" components of resistance make their noise contribution; but the fact remains that it is supported by careful experiment.

And now we come to reactance. Of itself it doesn't cause any noise, but it can affect resistance noise in two ways. It always brings with it a certain amount of resistance, corresponding to power losses; for instance, eddy currents in the wire of an inductor (and in the core, if there is one). And even a pure reactance affects the a.c. resistance of a circuit measured between any two points.

Calculating Noise Voltage

Fig. 1 is a simple example; in fact, the simplest that is anything like real life, for even an isolated resistor has some stray capacitance, which can approximately be represented by C, and which in practice would be augmented by the input capacitance of the amplifier without which the noise would not be noticed. How does one calculate the noise voltage between the terminals?

The noise formula, once again is

$$E = \sqrt{4kTBR}$$

$$\text{or } E^2 = 4kTBR$$

E being the e.m.f. of an imaginary noise generator in series with the resistance R. Of the factors involved, k is a known constant, 1.38×10^{-23} in the usual system of units; T is the temperature in degrees absolute or "Kelvin" (°K), and for ordinary room temperature can usually be taken as 290; and R in this case is R_p—or is it? And what is the frequency bandwidth?

There are several alternative ways of dealing with this problem. According to one point of view, sticking strictly to Fig. 1, the resistance is a fixed quantity R_p, which contains within itself the equivalent noise e.m.f. E; this is loaded by C, so there is a voltage drop in R_p depending on the reactance of C and therefore on frequency. That is where B comes in.

According to another point of view the voltage drop can apparently be dodged, because the impedance of R_p and C measured between the terminals can be exactly imitated at any given frequency by R_s and C′ in Fig. 2, provided that the values of R_s and C′ are correctly chosen for that frequency. The noise e.m.f. in R_s supplies no external current, so appears in full at the terminals, and there is therefore no need to allow for the capacitive loading. But the value of R_s depends on frequency, so in dodging one complication one runs into another. As a matter of fact it is exactly the same complication!

Comparison of Methods

Let us compare the two methods by lumping $\sqrt{(4kTB)}$ together and calling it K for short, and calculating the terminal voltage (call it V) by each method in turn.

Call the noise e.m.f. in Fig. 1 E_p, and the reactance of C X_c. Then

$$E_p = K\sqrt{R_p}$$

$$\text{and } V = E_p \frac{X_c}{\sqrt{X_c^2 + R_p^2}}$$

$$\text{So } V = KX_c\sqrt{\frac{R_p}{X_c^2 + R_p^2}}$$

Now call the noise e.m.f. in Fig. 2 E_s. Then

$$E_s = K\sqrt{R_s}$$

The formula for finding the value of R_s equivalent to R_p is

$$R_s = \frac{X_c^2 R_p}{X_c^2 + R_p^2}$$

$$\text{So } V = E_s = KX_c\sqrt{\frac{R_p}{X_c^2 + R_p^2}}$$

which is exactly the same as by the other method. So it doesn't matter whether we reckon the noise e.m.f. for the actual resistances in a circuit and then find the voltage between any two points by taking into account the rest of the circuit, or calculate or measure the equivalent resistance between the two points and apply the Nyquist formula directly to it.

Having disposed of that, we can tackle the bandwidth problem. According to the Fig. 2 point of view the bandwidth goes all the way from zero to infinity—or at least 6 MMc/s—but the equivalent resistance, and hence the noise voltage, falls with frequency, so the contributions at very high frequencies are negligible. According to the Fig. 1 point of view the resistance is definitely R_p but the proportion of noise voltage that actually appears between the terminals falls with frequency, so the bandwidth can be regarded as limited. As we have seen, both come to the same thing. If we use the Nyquist formula to calculate the terminal voltage squared *per cycle of B* at various frequencies, we can plot a graph like Fig. 3. The proper way of calculating the total noise voltage is to add together all these voltages-squared for every single cycle from zero frequency to infinity and take the square root of the result—or to be perfectly correct we should narrow the slices of bandwidth from whole cycles per second to infinitely small fractions and add together the infinitely large number of them.

To some readers such a task might seem to be in the same class as those set by unbelievably beautiful princesses in fairy stories in order to liquidate tiresome suitors. But, just as in the fairy stories, there is a magic formula provided by a powerful wizard, named Integral Calculus. Those who are not on good terms with this wizard will no doubt be obliged to make as good a guess as they can. They might pick out the frequency where the voltage-squared per c/s had fallen to a half, and reckon as if it existed at full strength up to that frequency (call if $f_{\frac{1}{2}}$) and then dropped to zero. In other words, they would substitute the shaded area in Fig. 3 for the area under the curve, reckoning R in the formula as R_p, and B as $f_{\frac{1}{2}}$, and trusting to luck that leaving out all the bits of noise beyond $f_{\frac{1}{2}}$ would just about cancel out the error of reckoning them in full up to $f_{\frac{1}{2}}$.

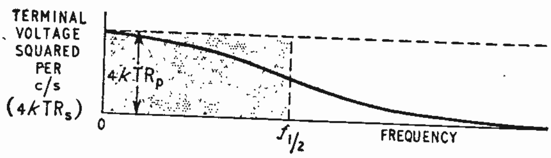


Fig. 3. Graph of the square of the noise voltage per c/s between the terminals of Fig. 1 plotted against frequency. Total noise is represented by the area under the curve; an approximation to it is the shaded area, the boundary $f_{\frac{1}{2}}$ being the frequency at which the curve is half its maximum height.

If the voltage-squared has dropped to a half at $f_{\frac{1}{2}}$, the voltage must have dropped to $\sqrt{\frac{1}{2}}$ or 0.707. Either way, it is a drop of 3 dB. This may remind us of the common practice of regarding the frequency band of a single resistance-reactance combination as ending where the loss is 3 dB. Besides "looking about right," this practice is convenient because the frequency chosen is the one at which the reactance is equal to the resistance; i.e.

$$\frac{1}{2\pi f_{\frac{1}{2}}C} = R_p$$

so

$$f_{\frac{1}{2}} = \frac{1}{2\pi C R_p}$$

But is this the $f_{\frac{1}{2}}$ we want? Is the frequency at which the impedance of Fig. 1 to signals is $\sqrt{\frac{1}{2}}$ times R_p necessarily the frequency at which the noise voltage is reduced in that proportion? If so, then R_s at that frequency must be $\frac{1}{2}R_p$. Well, one has only to equate $\frac{1}{2}R_p$ to the formula for R_s in terms of R_p to assure ourselves that it is. So we can substitute our formula for $f_{\frac{1}{2}}$ in place of B in the Nyquist noise formula, to discover that the terminal noise voltage works out at

$$\sqrt{\frac{2kT}{\pi C}}$$

The interesting thing about this is that resistance has completely disappeared, and so has frequency. Before we get too excited about this, however, we should remember it is only a guess, and wait to see the result of the consultation with Integral Calculus. This consultation is reported in full as Appendix I, and the result is remarkably simple:

$$\sqrt{\frac{kT}{C}}$$

The only thing wrong about the guess, then, is that it was $\sqrt{(2/\pi)}$ or 0.8 times the correct value, or 20% low. In practice, however, it might well be more nearly right, because even the widest-band amplifier doesn't go up to infinite frequency, so some of the noise included in the "true" formula is bound to be cut off. Of course, if the amplifier bandwidth is even less, say, than $f_{\frac{1}{2}}$, then one should use the basic noise formula, reckoning the amplifier bandwidth as B, and R_p as the resistance (if it is reasonably constant over the band).

The $V^2 = kT/C$ result is so simple and convenient, getting rid as it does at one stroke of both the rather hazy factors in the Nyquist formula, that one would expect it to be included in all the books and articles on the subject. It may be that I have looked up the wrong ones, but the nearest any of them go to it is to state the integral equation without bothering to work it out. No doubt innumerable irate authors will now write to point out that I have inexcusably overlooked their clear presentation of the matter. However, what I have overlooked others may have too, so it may help to press the point home if we see what sort of noise voltage to expect across a resistor—any resistor—at the input of a valve amplifier. The factor k is absolutely fixed, and for ordinary situations T is virtually fixed at about 290° K, while C is likely to be of the order of 10pF. So

$$V = \sqrt{\frac{1.38 \times 10^{-23} \times 290}{10^{-11}}} = 20 \times 10^{-6}$$

$$= 20 \text{ microvolts.}$$

It is seldom likely to be much more than this, because to double it would necessitate reducing C to 2.5pF, and almost any valve's C_{in} would contribute more than that. It might be much less, however, and this formula is useful for estimating the noise from a crystal or electrostatic microphone. But except for high resistances the noise voltage actually producing results at the output would usually be less—perhaps much less—than in this formula because of the restricted bandwidth of the amplifier. So it is useful to have some idea of how much of the theoretically infinite bandwidth represented in the kT/C formula does actually contribute appreciable noise. And this is almost where we came in, for we have already done something like it before—with the help of Fig. 3. But we can do it rather more scientifically now by taking as the effective bandwidth of the Fig. 1 circuit the bandwidth that would include the same amount of noise from a constant resistance equal to R_p. This can easily be found by equating the noise-squareds given by the two formulae:

$$\frac{kT}{C} = 4kTR_p B$$

Whence

$$B = \frac{1}{4CR_p}$$

Since B begins at zero this really gives the "effective" top frequency; call it f_e. (An alternative way of getting at it is from what we already know, that it is $\pi/2$ times $f_{\frac{1}{2}}$.) Here, from the use of this result, are some representative values of f_e on the assumption that C = 10pF (see table, left).

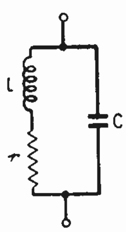
R_p	f_e
1 Ω	25 kMc/s
10 „	2.5 „
100 „	250 Mc/s
1 kΩ	25 „
10 „	2.5 „
100 „	250 kc/s
1 MΩ	25 „
10 „	2.5 „

So for the lowest values of resistance the noise voltage indicated by the kT/C formula has no great practical significance, because no amplifier would have sufficient bandwidth to make all of it evident, and even if it could its own noise would swamp the resistance's. Even the medium values are likely to be limited more by the amplifier than by C. But with resistance of the megohm order the effective noise voltage may well be as much as 20μV.

A still more important noise maker is the ordinary tuned circuit, which can be represented by Fig. 4. Its resistance, measured between the terminals, increases from the relatively small value r at zero frequency to a maximum R_d at resonance and then down almost to nothing at very high frequencies. (Incidentally it should be remembered that r itself varies somewhat with frequency.) This is a good example of how the apparent resistance rather than the actual resistance in the circuit counts for noise voltage; for the smaller r is, the larger R_d. In this particular case it is easy to see why the noise voltage should correspond to R_d rather than r, because those parts of the noise voltage generated in r at or near the resonant frequency are magnified by the resonance effect, just as would signals injected into the circuit at the same place.

V.H.F. Tuned Circuit

On presenting this problem to our wizard, we receive from him some rather awkward-looking instructions, and the usual procedure is to by-pass these by arguing that an amplifier with a tuned input circuit will be almost certain to have a more or less clearly defined frequency band no wider than that over which the input circuit dynamic resistance is at or near its maximum value, R_d. So the Nyquist formula is used, with R_d as R and the amplifier bandwidth as B. To get an idea of the sort of values, let us consider a v.h.f. circuit with a dynamic resistance of 10kΩ followed by an amplifier with a bandwidth of 500kc/s. Then $\sqrt{(4k\,TBR)} = \sqrt{(4 \times 1.38 \times 10^{-23} \times 290 \times 5 \times 10^5 \times 10^4)} = 9\mu V$. Typical v.h.f. single tuned circuits with $R_d = 10k\Omega$ would generate appreciable noise over a wider band than 500kc/s; so in this case the amplifier selectivity is cutting the noise somewhat. This is just as well when calculating by Nyquist; because it ensures that the resistance is not too far below R_d at the edges of the frequency band.



(Right) Fig. 4. Another important circuit—the parallel tuned circuit.

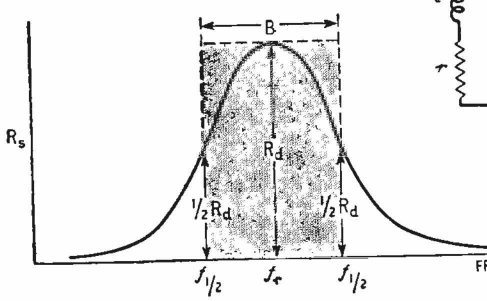


Fig. 5. This diagram can be used in estimating the total noise voltage between the terminals in Fig. 4.

However, if we want to have an idea of the noise voltage at the terminals of the circuit itself, irrespective of how it may be pruned by the selectivity of the amplifier, we can approximate to it by the method we used for Fig. 2 in Fig. 3. Of course the curve falls away in both directions from the resonant frequency, which we will call f_r, as in Fig. 5. So there are two frequencies at which the noise voltage-squared (and therefore the equivalent series resistance R_s) is half the maximum (R_d), and the distance between them is what we will regard as B.

Now it can be shown (see Appendix 2) that, as in Fig. 1, these $f_{\frac{1}{2}}$ points are also the frequencies where a signal would be reduced by 3dB as compared with spot-on tuning. And it is well known that the frequency band B between these points is very nearly 1/Q times f_r. And as $Q = 2\pi f_r L/r$ we can calculate B thus:

$$B = \frac{f_r}{Q} = \frac{f_r r}{2\pi f_r L} = \frac{r}{2\pi L}$$

Then we fill this value for B into Nyquist:

$$V = \sqrt{4kTBR}$$
$$= \sqrt{\frac{4kTrR}{2\pi L}}$$

We are assuming that R has the value R_d over the band B, and R_d is known to be L/rC, so putting this in we get

$$V = \sqrt{\frac{4kTrL}{2\pi LrC}}$$
$$= \sqrt{\frac{2}{\pi} \cdot \frac{kT}{C}}$$

This has a strangely familiar look—yes! it is identical with the result we obtained for Fig. 1 by the same method, and which we found to be of very much the same order as the theoretically exact figure $\sqrt{(kT/C)}$ obtained by integration. So it should not be surprising to learn that this is the theoretically exact figure for the tuned circuit too.

Again, the value of this delightful simplicity tends to be rather lost because so often the amplifier and what have you restricts the frequency band that applies so far as results are concerned. And then Nyquist may be more helpful.

Other Resistors?

Mention of the amplifier may lead to a question on why I have been assuming all along that the noise comes from one particular resistor—the one at the input of the amplifier—as if all resistances in the amplifier or indeed anywhere were not noise-makers. Well, of course they are, but the point is that their noise is usually amplified so much less that it can be neglected without making much difference. A possible exception would be the coupling between first and second stages, if the first had an unusually small amplification, as might be so at very high frequencies. A more pointed objection to any assumption would be that the first valve itself contributes noise, which is rarely negligible. It is "shot noise," which is similar in some ways to thermal noise and different in others. There is no space here to go into these differences, except to mention that the noise current is generally proportional to the square root of anode current. In practice, first-valve noise is usually lumped in with first-circuit noise by specifying it as "equivalent noise resistance"; that is to say, the resistance which, if connected at the input of an imaginary noiseless but otherwise equal valve, would cause the same noise at the output. So if the equivalent noise resistance of a valve were 3,500 Ω and the input circuit had a resistance of 8,000 Ω, the noise output would be the same as that given by a noiseless valve with a 11,500 Ω input circuit.

Lastly, if the noisiness of a given resistor depends only on the things already mentioned—temperature, *k*, etc.—and not at all on the quality of resistor, why are some classed as "low noise"? Is it just advertisers' licence, without any scientific basis?

By no means. The thermal noise we have been discussing at such length is only one of the kinds in non-metallic resistors, which are by far the most commonly used. In carbon composition resistors it may contribute only a small fraction of the total noise. The reason it has been (I hope) worth discussing is that it is an irreducible minimum, so it forms a standard by which total resistor noise can be judged. The other noise is caused by variations in resistance when current is flowing, and these of course result in fluctuations of voltage across the resistor. Unlike thermal fluctuations, they are worst at lowest frequencies; they tend to predominate over thermal noise below about 10kc/s. As one would expect, they are proportional to current. In "high-stability" cracked-carbon-film resistors the current noise is very much lower, being almost unmeasurably low except for low and medium resistances but rising rapidly above 1 MΩ. One should of course never use a composition resistor across the input of a high-gain amplifier.

APPENDIX I

Calculation of the noise voltage V at the terminals of R_p and C in parallel (Fig. 1) over an infinite frequency band.

$$V^2 = 4kT\int_0^\infty R_s df \qquad \text{where } R_s = \frac{R_p}{(\omega CR_p)^2 + 1}$$

$$= 4kTR_p\int_0^\infty \frac{df}{(2\pi fCR_p)^2 + 1}$$

$$= \frac{4kTR_p}{(2\pi CR_p)^2}\int_0^\infty \frac{df}{f^2 + \frac{1}{(2\pi CR_p)^2}}$$

$$= \frac{2kT}{\pi C}\Big[\tan^{-1} 2\pi fCR_p\Big]_0^\infty$$

$$= \frac{2kT}{\pi C}\cdot\frac{\pi}{2}$$

$$= \frac{kT}{C}$$

$$\text{So } V = \sqrt{\frac{kT}{C}}$$

APPENDIX II

The series resistive component R_s of the impedance of the Fig. 4 circuit between the terminals is

$$R_s = \frac{r}{(1 - \omega^2LC)^2 + \omega^2C^2r^2}$$

Let $R_{\frac{1}{2}}$ be the value of R_s at the points on the slope of the resonance curve where the whole impedance is $\sqrt{\frac{1}{2}}$ times that at resonance (R_d). Unless resonance is very flat, the ratio of the resonant frequency, f_r, to the frequency difference between these points, B, is very nearly equal to the Q of the circuit. With the same proviso, $1 - \omega^2LC$ at either of these points is very nearly equal to B/f_r.

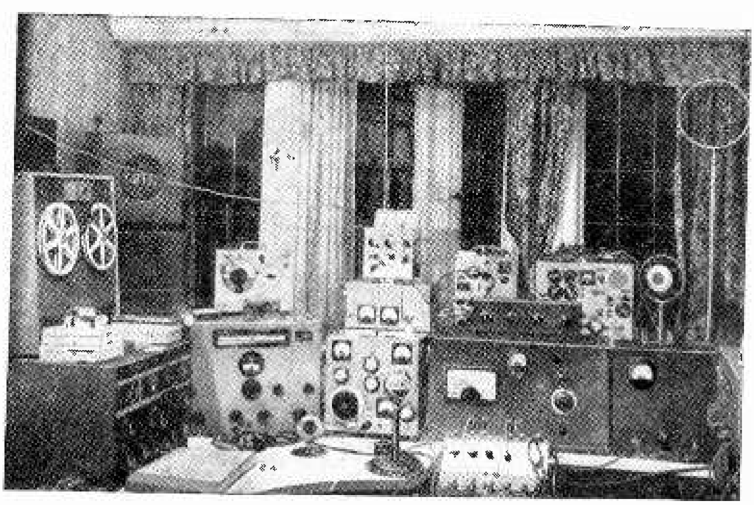
$$\text{So } R_{\frac{1}{2}} \simeq \frac{r}{\frac{1}{Q^2} + \frac{1}{Q^2}} = \frac{rQ^2}{2}$$

At resonance, $1 - \omega^2LC = 0$, so

$$R_d = rQ^2$$

$$\text{So } R_{\frac{1}{2}} \simeq \frac{R_d}{2}$$

Mr. Wiktor Tell has pointed out a slip in the algebra at the end of the May instalment. Instead of saying v is proportional to T it should have been $\sqrt{T}$. The final conclusion then works out to $I \propto \sqrt{T/R}$, which really *is* in agreement with previous results!

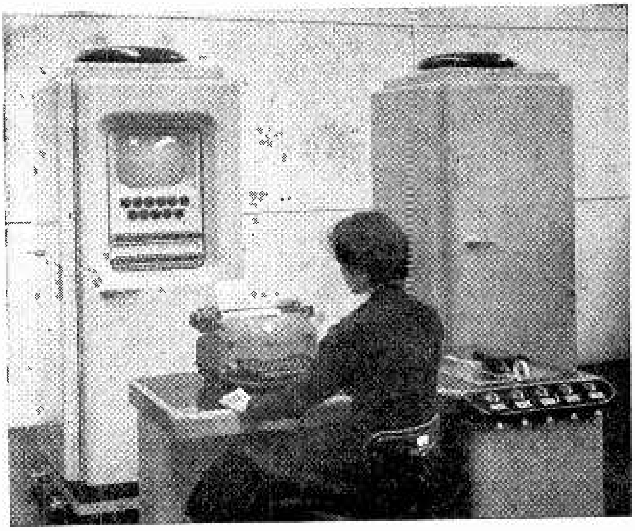


C. H. L. EDWARDS, author of the article in our March issue describing an amateur transmitter-receiver, is well known in amateur circles as operator of G8TL, at Theydon Bois, Essex. On the left in this photograph of the station are two receivers (AR88 and CR100), in the centre is a 160-m transmitter to the right of which is an all-band 75-watt transmitter and a 35-watt modulator. The transmitter-receiver described in the March issue is the smaller of the two sets behind the main transmitter.

Physical Society's Exhibition

NEW ELECTRONIC DEVICES AND TECHNIQUES

Valve and allied devices as well as test and measuring instruments shown at this exhibition are described in a separate report in this issue.



Decca Radar digital computer using magnetic-core circuits.

RESEARCH

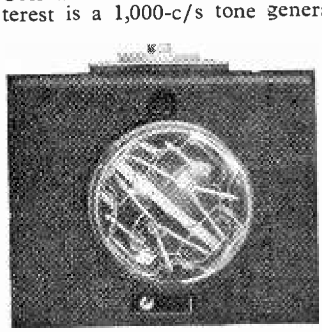
Communications.—The assessment of multi-channel communication systems from the point of view of noise and intermodulation is a tedious process when carried out by normal single frequency measurements. Quicker and more realistic results are obtained in a method developed by G.E.C. Research Laboratories in which the total multi-channel signal is simulated by noise of the appropriate bandwidth. If band-stop filters are introduced at the source to represent unoccupied channels and the output in these channels is examined after passing through the system and related to the signal power in an "occupied" channel of equal bandwidth, the ratio enables a rapid assessment of performance to be made.

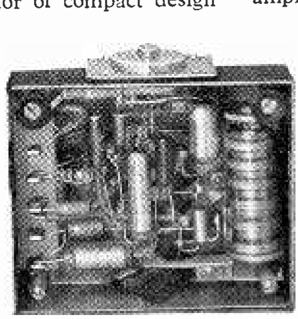
A time-division method of transmitting two-way audio signals on a line without the use of amplifiers was shown by the Post Office Research Station. The system has been developed for use with electronic exchanges where the power required for switching is small compared with conventional electromagnetic relays. The energy stored in the terminal capacitors of filters at each end while the line is open is transmitted in pulses when the line is closed by applying a 3-μsec pulse at 100-μsec intervals to a biased diode switching system. If there is no dissipation in the filter components all the energy is concentrated in the pulses, and in practice useful distances can be covered without any amplification.

Oscillators.—Another Post Office development of interest is a 1,000-c/s tone generator of compact design for use as a frequency standard in the field. The resonator is a gapped quartz ring about 1 inch in diameter which is maintained in vibration by a transistor amplifier. Sound is emitted by a hearing-aid earpiece, loaded by a quarter-wave pipe. In the region of 20° C the temperature coefficient is 5 parts in 10^7 per degree C.

For the maintenance of oscillations of very low frequency (less than 1 c/s) thermistors may be used to simulate inductance by virtue of the thermal lag of current behind the applied voltage. The Radio Research Station showed a simple parallel capacitor-thermistor combination with a period of several seconds working in conjunction with a pen recorder. It was stated that the equivalent inductance of a type A5412/100 thermistor under these conditions was 8,550 henrys!

Semi-conductors.—Research into semi-conductor alloys, and in particular indium antimonide, has been extended by the Services Electronics Laboratory. The electron mobility in this material is 17 times that of germanium and the Hall effect, for a given transverse magnetic field, gives an increase of available power at the side electrodes, proportional to the square of the mobility, of about 300. With Mumetal rods used as flux concentrators the device makes either a sensitive electrical compass or a "gradiometer" for detecting anomalies in the earth's field in magnetic survey work. A deviation of one degree from the null position, at right angles to the field, gives a power output of 0.002 μW, which will operate a galvanometer or sensitive relay without amplification.





Portable 1,000-c/s tone generator with gapped quartz ring vibrator (G.P.O.)

ELECTRONICS

Computing.—Decca Radar have now entered this field with a digital computer which is notable for its small size, reduced power consumption and heat dissipation and low cost. These features are the direct result of using two-state magnetic cores instead of valves, not only for storage purposes but throughout the arithmetic circuits as well. Valves are used merely for generating pulses for driving the core circuits. The equipment is constructed on the "package" principle, with the 2-mm diameter cores mounted on printed circuit cards which plug into racks. The main store is a magnetic drum and this rotates at a speed (6,000 r.p.m.) which makes its digit rate three times that of the computing circuits, so that the normally long access time is reduced to a third. For scientific work the input and output medium is normally punched paper tape, but for business applications magnetic tape equipment is available as well and this can also serve as long-term storage for reference data.

Representative of electronic analogue computing was a "real-time" simulator for the study of control systems, shown by Elliott Brothers. Designed on the unit construction principle, it was actually built up from three of the Elliott general-purpose machines described last year but had additional apparatus including a non-linear function generator and an electronic multiplier of the crossed-fields type.

Storage Systems.—A new type of two-state storage device based on the non-linear dielectrical properties of barium titanate was demonstrated on the Plessey stand. This material has an electric field-strength/flux-density characteristic which takes the form of a square hysteresis loop, giving two states of remanent induced charge. The storage cell itself consists of a single crystal of barium titanate with a small matrix system of electrodes on either side of it (X co-ordinate electrodes on one side and Y on the other), and application of a voltage pulse (say 40 V) to one X electrode and one Y electrode induces a charge at the point of intersection. To "read out" the stored information a higher voltage pulse (100 V) is applied. If this is of the same polarity as the induced charge it produces no output, but if it is of opposite polarity the reversal causes an output pulse to appear on the appropriate co-ordinate electrodes. The cell has the advantage over magnetic two-state stores of smaller size and lower power consumption, but requires higher voltages. A "read-out" speed of 0.2 μsec can be obtained.

The use of ferrite magnetic cores for binary storage devices, using the square hysteresis characteristic, is now well known, and examples of matrix stores based on this principle were shown by both Plessey and Mullard. The Mullard magnetic matrix is particularly interesting for the use of printed circuit frames for terminating the threaded wires, giving a neat and simple form of construction.

Magnetic tape is one of the most obvious media for storing information but has not hitherto been used a great deal in this country. However, Ferranti were showing a new magnetic tape handling equipment for computers, and this was notable for its high tape speed of 100 inches per second and its ability to start and stop within 10 milliseconds. Four channels can be accommodated across the width of the ½-inch tape and in each channel the digit handling rate is 10,000 per second.

Transistor Instruments.—The use of transistors in oscillator circuits to give small and compact e.h.t. supplies was exemplified by two instruments shown by the Atomic Weapons Research Establishment—a gamma radiation monitor and a charging unit for quartz-fibre radiation dosimeters. Both are powered by two 1.35-V cells and are small enough to be carried in the pocket. The portability offered by transistors is also of great value in physiological amplifiers, enabling them to be brought nearer to the patient to avoid pick-up of hum and other interference. Here, examples were shown by Edison Swan and Guy's Hospital Medical School.

In the field of time measurement, B.T.-H. had a transistor timebase calibrator which can measure intervals from 10 μsec to 2,000 μsec with an accuracy of 0.5 μsec, while Mullard demonstrated a transistor counter-chronometer which can also be used for frequency measurement. In the last-mentioned instrument a quartz-crystal oscillator and transistor frequency dividers produce reference timing pulses with accurately known repetition rates from 100,000 to 1 per second. A time interval to be measured (between two input pulses) is then determined by counting the number of reference pulses which occur during the period. Alternatively, frequency measurement is performed by counting pulses derived from the input waveform during the period of 1 second.

Machine Tool Control.—A system for setting the co-ordinates of a machine tool work table was demonstrated by B.T.-H. The scale for each motion consists of a composite bar with magnetic and non-magnetic segments with interfaces spaced exactly 1 inch apart. As the table moves the bar varies the flux in a differential pickup head, any difference in the two paths being amplified and used to control the servo motor driving the lead screw. When the flux is balanced a movement of 0.00002in causes a detectable error signal. Intermediate dimensions are registered by a subsidiary micrometer screw servo drive which alters the setting of the pickup head.

Ultrasonics.—Specialized echo sounding equipment for locating the bearing and range of whales has been developed by Kelvin Hughes. The horizontal scanning beam can be varied in vertical width and alternative frequencies and pulse lengths are provided to give optimum conditions of detection from 1,800 metres down to the firing range. Both aural and visual indication of the returning echoes is available.

Ultrasonic abrasion (drilling) of hard materials has been carried a stage further by the development by Mullard and Plessey of small portable drills. The Plessey drill uses a new piezoelectric ceramic material ("Casonic III") with enhanced power-handling capacity and frequency stability.

Equipment for the ultrasonic cleaning of small parts on a laboratory scale has been developed by Dawe Instruments. To ensure effective cavitation in the cleaning liquid at the surface of the objects, a pulsed oscillator is employed which also operates with less power.

An ultrasonic thickness gauge designed to work with a 50-ft connecting cable has also been introduced by Dawe. It is designed for us in shipyards where it is often inconvenient or hazardous to take the main equipment close to the part of the hull to be examined. The equipment works on the resonance principle and the frequency is adjusted until the wavelength in the material is a function of the thickness.

The calibration of barium titanate accelerometers at high frequencies and small amplitudes (of the order of 10^{-6} inch) calls for special methods of measuring the amplitude, and E.M.I. Electronics demonstrated an optical interference principle which has been found useful in this application. An interferometer of the Michelson type is mounted above the vibration table to which one of the mirrors is fixed. With monochromatic light interference fringes are visible in the interferometer eyepiece, and as the amplitude is slowly increased these alternately disappear and reappear. Each disappearance is sharply defined and is a subjective phenomenon based on the time average of the instantaneous brightness. This can be calculated, and related to the amplitude and wavelength of the light.

Photographic.—In photographic printing the range of densities in the negative is often too great for the printing paper to reproduce all the detail in the picture. To overcome this Cinema-Television have introduced an electronic contact printer which smooths out such large variations by a feedback principle. The negative, with printing paper behind, is scanned by a spot of light from a c.r. tube in television fashion, with overlapping of lines to give complete and even illumination. A photocell then picks up the transmitted light on the distant side of the printing paper and gives an output signal which varies with the density of the negative. This signal is amplified and fed back to control the brightness of the c.r. tube scanning spot so that it decreases in intensity when the less dense areas of the negative are scanned. In this way the effects of the density variations can be reduced to any desired extent by controlling the degree of feedback.

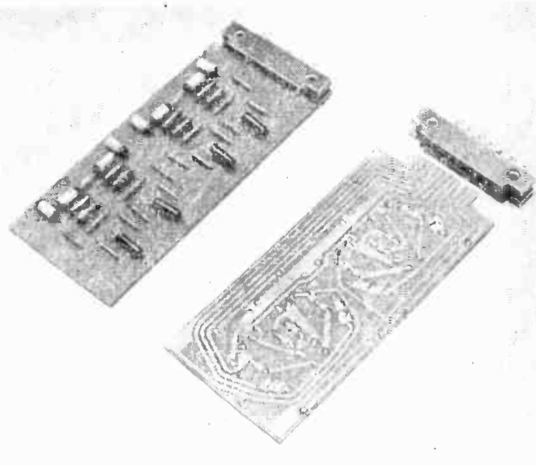
Scanning of a kind is also used in a new method of high-speed electronic photography developed by Mullard. This gives a sequence of very short exposure pictures and uses a device similar to an image converter called an image dissector tube. The photocathode of the tube, which is exposed to the phenomenon to be photographed, is active only in patterned areas, and the electrons from these are swept across the fluorescent screen by a deflection system. This results in a composite picture on the screen which is recorded by a camera. The individual pictures from the composite record can be sorted out afterwards by viewing through a patterned transparency or by means of another dissector tube. Recording speeds of up to 10 pictures per millimicrosecond are possible.

Displacement.—A convenient method of measuring displacement has been evolved in the Mechanical Engineering Research Laboratory of D.S.I.R. A Ferroxcube core attached to the moving part (e.g., a valve tappet) is arranged simultaneously to increase and decrease the inductance of a pair of coils forming two arms of an inductance bridge. The other pair of arms is formed by a similar transducer in which the core is adjustable by a micrometer. The bridge is excited at 10 kc/s and any out-of-balance component is detected by a one-cycle response demodulator* and indicated on a centre-zero meter. Alternatively, the balance point can be used to brighten the trace on a c.r.t. display and the amplitude measured at any point by reference to the micrometer reading.

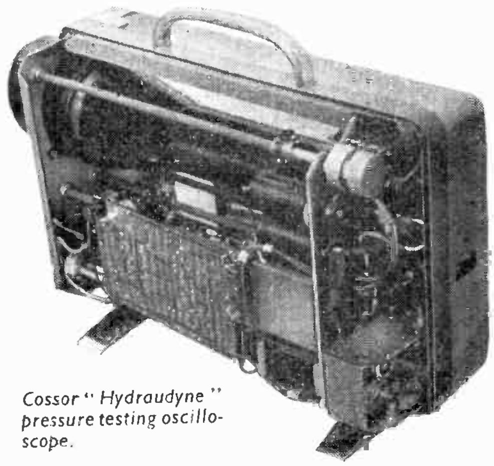
Volume.—The principle of the Helmholtz resonator, familiar in connection with loudspeaker cabinets, is used in a "Volumometer" which was shown by the research department of the Morgan Crucible Company. The change in resonant frequency resulting from the introduction of a body into the resonator chamber depends only on its volume and the reduction of compliance of the remaining enclosed air, and is independent of the shape.

Pressure.—For investigations and routine testing of hydraulic transmission systems over a range of pressures from 25 to 50,000 lb/in^2 Cossor Instruments have in-

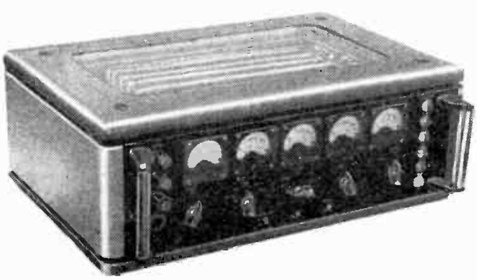
* E. A. Johnson. Selected Government Research Reports, Vol. 5, Report 9, pp. 128-133. (H.M. Stationery Office.)



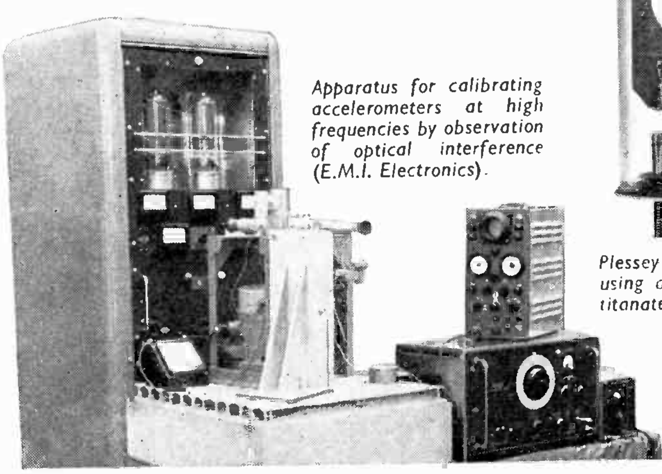
Ferranti transistor digital computing "package" using printed circuit.



Cossor "Hydraudyne" pressure testing oscilloscope.



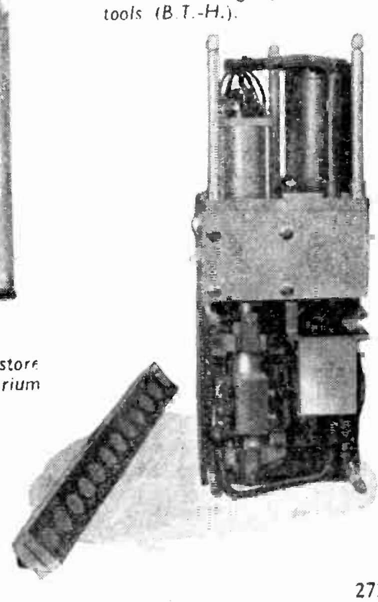
Mullard transistor counter-chronometer.



Apparatus for calibrating accelerometers at high frequencies by observation of optical interference (E.M.I. Electronics).

Plessey matrix store using a single barium titanate crystal

Below :—Measuring bar and electro-magnetic head for co-ordinate setting of machine tools (B.T.-H.).



troduced the "Hydraudyne," which comprises a pressure transducer, amplifier and c.r. oscilloscope with a long-persistence screen. The timebase has a scanning velocity variable from 0.001 to 10 sec and the amplifier response has a rise time of 10 μsec.

The radial pressure built up on textile thread bobbins by successive layers can be measured by a transducer developed by the Rayon Research Association and made by H. Tinsley (Type 30-A). A wire resistance strain gauge records the transverse distension of a pad of resilient material which is inserted under the layers of thread.

Flow.—A flowmeter (Type ND.31) for conducting liquids, developed by Elliott Brothers, works on an inversion of the electromagnetic principle used for pumping liquid metals. An alternating magnetic field traverses the flow tube at right angles to a pair of diametrical probe electrodes, and any movement of the liquid induces an e.m.f. which can be amplified and applied to a meter, recorder or servo controller. The indication is independent of conductivity over a wide range (tap water was used in the demonstration) and an accuracy of ±1 per cent of full scale is claimed over the range 0-400 cm^3/minute. Originally, the development was carried out in conjunction with A.E.R.E., Harwell, for measuring rates of flow of radioactive nitric acid.

MATERIALS

Microwave Absorption.—Sheet material (AF 11) consisting of prepared foam rubber with absorption between 99 per cent and 99.7 per cent in the range 5 to 35 kMc/s has been introduced by Plessey. It is available in panels 1ft square and is suitable for lining laboratories where field measurements are to be made at these frequencies. Another range ("M" type) depends on an interference principle in which the energy reflected from a metal backing is absorbed in facing material of critical magnetic and dielectric properties and high refractive index; the total thickness is considerably less than the free-space wavelength of the radiation.

Magnetostrictive Ferrites.—The use of these materials as an alternative to electrostrictive ceramics such as barium titanate is based on the fact that lower operating voltages are required. They are also superior to metals such as nickel in having low conductivity and requiring no lamination to reduce eddy currents. Plessey as well as Mullard are actively engaged in developing these materials.

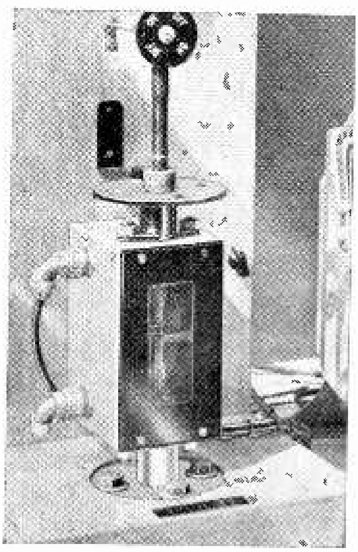
Magnetic Tape.—As an alternative to magnetic iron oxide, *pure* iron in powdered form is being used as a coating medium in ¼in-wide tapes made by Salford Electrical Instruments. When the particle size is sufficiently reduced pure iron (in bulk a "soft" magnetic material) acquires "hard" magnetic properties and exhibits a high remanence. The tapes shown have a sensitivity 6 dB below normal oxide tapes but 15 dB higher maximum output and, it is claimed, lower harmonic distortion. They can be made in a wider range of coercivities and they cause less abrasion of the recording heads than oxide types.

Laminations.—The production of awkward shapes or small quantities for which the cost of press tools would be prohibitive is facilitated by an etching process which has been developed by the Telegraph Construction and Maintenance Company. The process is similar to that used for printed circuits and can be used for laminations up to 0.004in thick.

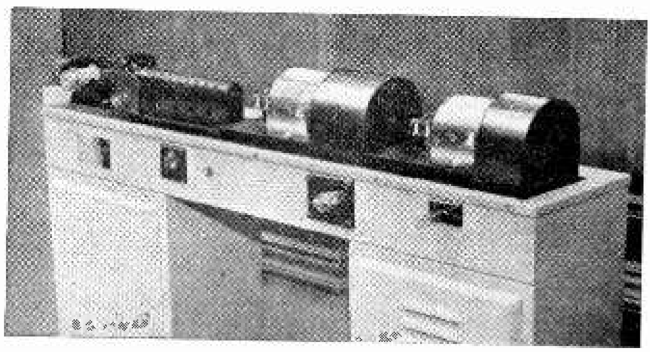
Miscellaneous Exhibits.—Sub-miniaturization continues to be one of the main trends in component development, and most of the newest types exhibit this feature. Plessey have extended their range of "Castanet" tantalum electrolytic capacitors and added some new models primarily for transistor applications. These are housed in metal cases only ¼in in diameter and under ½in long. A 45-μF, 10-V type, for example, is 7/16in long.

The same trend was observed in the new Type SD wirewound potentiometer shown by Salford. It is a semi-precision type in a metal case, yet measures only ⅝in in diameter and ½in deep, and at present is available in resistance values of from 2 kΩ to 8 kΩ. The rating is 0.5 W.

Two very small moving-coil relays were seen this year; one made by Elliott is argon-filled and hermetically sealed and plugs into a B9A valveholder. It operates with only 10 mW input and its contacts will handle up to 2 W. The other is housed in a small rectangular case with flying leads and is made by Electro Methods. It, also, operates with only 10 mW



Electromagnetic flowmeter (Elliott Brothers).



Above: Miniature "cable-laying" machine shown by Fortiphone.

Left: Salford sub-miniature precision potentiometers, Type SD.

input and handles 2 W. Despite the sensitivity of these units they are said to be particularly robust and resistant to shock and acceleration.

An aid to the further miniaturization of mobile v.h.f. radio-telephone equipment is the extension of the overtone operation of quartz crystals into the regions of 100 and 200 Mc/s. Crystals operating on their 7th overtone (180 Mc/s) were shown by Standard Telephones and some 5th-overtone models (up to 75 Mc/s) by Salford. Apart from permitting a reduction in size of equipments these crystals lead to a worth-while economy in operation as in many cases no frequency multiplying stages are required. Reliable and economical crystal control of f.m. receivers becomes a possibility.

A new telephone receiver (earpiece) was shown by Standard Telephones in which higher sensitivity (about 5 dB above the average) is achieved by separating the magnetic and acoustic functions. The former is performed by a rocking armature and the latter by a lightweight metal cone-shaped diaphragm. In some respects it resembles the early balanced-armature loudspeaker movements, but is very much smaller and far more sensitive. The frequency response is linear to within ±3 dB from 200 to 3,500 c/s. A similar unit, with the frequency range extended to 4,000 c/s, is used as a transmitter (microphone).

Among the exhibits of Fortiphone was a miniature (for its type) "cable-laying" machine. Its function is to produce a twisted two-wire insulated cable using very fine wires. In one form it "lays" two strands of 0.0036in diameter (43 s.w.g.) heat-stripping enamelled copper wire with, initially, two strands of 0.005in diameter nylon monofilament and a final overlay of 7 strands of nylon. A plastic adhesive is automatically applied and a small oven dries it off before reeling. Activities in the ultra-miniaturization field have created a demand for such a cable, which was hitherto unobtainable anywhere. A four-way cable is provided by cementing together two twin cables. The breakdown voltage is around 5,000.

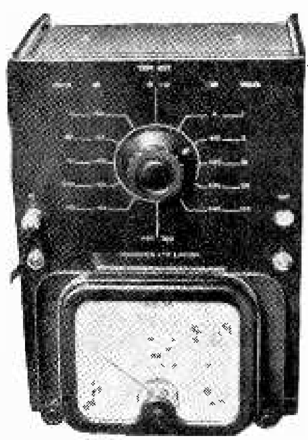
TEST AND MEASURING GEAR

New Exhibits at Recent Shows

This report embraces instruments shown at the R.E.C.M.F., Physical Society and 2nd International Instrument exhibitions without distinction. The last-named show introduces some unfamiliar marques. As in previous reports, many of the instruments mentioned were shown in prototype form and are subject to modification before they become available, if they do. Those previously reported are not mentioned again, even if now available for the first time, unless the modifications are substantial.

One design trend should be mentioned at the start, because it is not confined to any particular kind of instruments—the use of transistors. An increasing number of indicators employ them in place of valves for amplification, with great saving in space and power consumption. A number of transistor oscillators also appeared; crystal controlled and otherwise. Transistors are particularly appropriate for bridge oscillators and null indicators. Other transistorized instruments included a distortion-factor meter, a frequency meter, a time-base calibrator and a complete oscilloscope.

Circuit printing has not yet made much headway among measuring instruments, presumably because the production quantities are seldom large, but the Avo AM/FM signal generator and the Dawe sound-level meter have made a start with this technique, which for instrument work has an obvious advantage in reducing manufacturing variations.



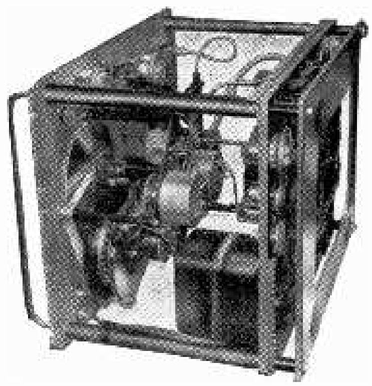
Right: Grayonics millivoltmeter, Type 1277.

Below: Pye "Scalamp" electrostatic voltmeter.

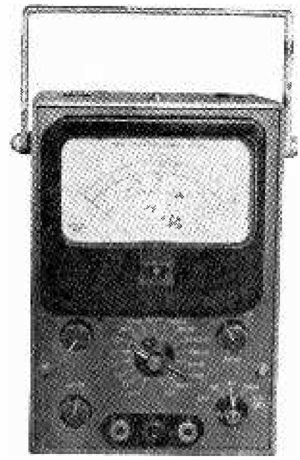
Developments in unamplified meters were less noticeable this year than last, but to the Pye "Scalamp" series of electrostatic voltmeters has been added an enlarged model for measuring the higher tensions used for cathode-ray tubes, up to 40 kV. At this level the prevention of brush discharge calls for special attention. The Labgear r.f. millivoltmeter, covering 20 mV to 32 V in six ranges, at frequencies from 50 kc/s to 250 Mc/s, might easily be mistaken for a valve-aided instrument, but actually employs only a germanium crystal rectifier operated on the slide-back principle, and its power needs are confined to a small bias battery.

The evolution of indicating instruments from the original valve voltmeter continues however; especially in the two directions of higher sensitivity and either wider or narrower frequency bands. The Grayonics millivoltmeter has no fewer than 12 linear voltage ranges, from 1 mV to 300 V full-scale, over the frequency range 20 c/s to 2 Mc/s. Its meter can be adjusted to the most convenient angle for reading.

For audio frequencies much higher sensitivities are possible; there is the B & K (Bruel and Kjaer; Denmark) Model 2408 with full-scale readings from 31.6 μV to 1,000V over the range 20 c/s to 20 kc/s. Although the lowest f.s.r. on the Peekel (Holland) Model 051B is higher—10 mV—this instrument is notable for its frequency range going down from 10 kc/s to 0.15 c/s, for which pointer instruments are not usually available. Three steps of pointer damping are provided. Even this is not quite zero frequency; for it the same firm lists Model 30B, with 10 ranges from 30 μV to 1V f.s.r. Zero drift is claimed as less than 2 μV per hour.

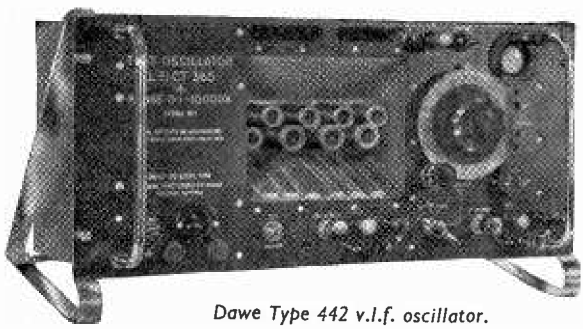


Frequency control gear in Sanders XT312 signal generator.



New London transistorized volt-ohmmeter.

Marconi Instruments alignment oscilloscope, Type TF1104.



Dawe Type 442 v.l.f. oscillator.

The B & K Model 2423 megohmmeter, 0.1 MΩ to 10 MMΩ, can also be used as a voltmeter or a micro-microammeter. Elliott make a micro-microammeter (which presumably can be used as a megohmmeter) covering 10 μμA to 1 μA f.s.r. in six ranges, using a ME1401 electrometer valve in a sealed and desiccated sub-unit. Further developments in the Pye galvanometer-modulator system of sensitive z.f. amplification enable a large number of low voltage and current ranges to be obtained.

The New London (U.S.A.) transistorized volt-ohmmeter has d.c. and a.c. (10-5,000 c/s) ranges with only 5 μA full-scale load; voltages 0.03 to 1,000V f.s.r. in 10 ranges, and resistances 1, 10, 100, 1,000 and 10,000 Ω mid-scale.

Versions of an a.f. wattmeter of N.P.L. design are offered by Cambridge Instrument and by Tinsley; the numerous ranges down to 2.5 mW full-scale on a conventional dynamometer instrument are made possible by negative-feedback amplifiers for both current and voltage coils. Another instrument, in which an amplifier is used to bring noise voltages up to the level at which a mean-square detector can be efficiently operated, is the Wayne-Kerr video noise-level meter. Its bandwidth is flat from 10 kc/s to 1.5, 3, 6 or 10 Mc/s according to the setting of a switch.

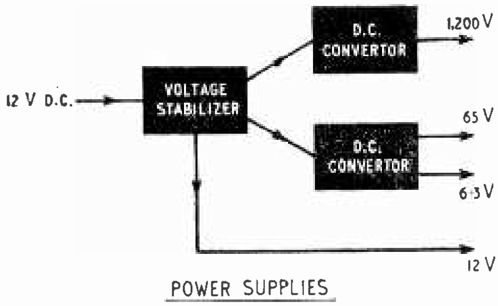
Frequency-selective amplification is used mainly for two purposes; to enable a sharp minimum to be obtained in bridge work; and for frequency analysis. The Microwave Instruments Type 3100, which would be very effective as a null indicator, was actually developed primarily for use with crystal detectors in microwave gear, and achieves the exceptional sensitivity of 1 μV f.s.r. clear of noise by the sharpness of its tuning to 1,025 c/s. A smaller instrument by the same firm uses OC71 junction transistors to give a sensitivity of 20 μV f.s.r. at 3,200 c/s, or slightly better on a flat range from 100 c/s to 10 kc/s. The B & K Model 2002 employs the superheterodyne principle to provide ranges down to 15 μV f.s.r. from 20 c/s to 30 Mc/s, the i.f. being 1,650 kc/s. A very different B & K selective indicator is their Model 2105, for a.f. analysis over 47 c/s to 12.5 kc/s in eight bands and with four steps of selectivity in a degenerative RC amplifier. Another B & K instrument that can be used for frequency analysis is their filter set comprising 27 one-third octave filters covering in all 40 c/s to 16 kc/s. A useful filter is the Krohn-Mite (U.S.A.) Model 310AB, a RC network with amplification, having the unusual facility of separate adjustment of low and high cut-off frequencies over the range 20 c/s to 200 kc/s; input impedance 6 MΩ in parallel with 50 pF, and output 500 Ω.

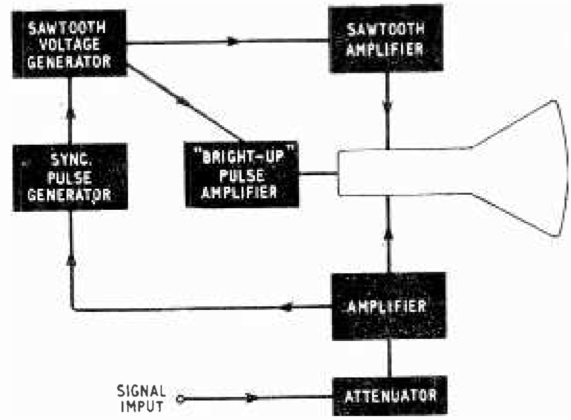
The new oscilloscopes have been designed for improved refinement of control or facilities, such as width of frequency band or ability to select small portions of complicated waveforms. Signal delay, so that non-repetitive waveforms can be observed in full, is now quite frequently provided, even in general-purpose models such as the Mullard L140 precision type. The dual trace L101 oscilloscope now appears with certain improvements as a Mark 2; for example, time-base triggering from either positive or negative wave fronts, and continuous sweep expansion. The latter facility, over a 10 to 1 range, is included in the Solartron AD557 with d.c. amplication for both X and Y. Nagard, who specialize in wide-band oscilloscopes, showed their DE103 with an improved specification, including signal delay line and frequency range from 0 to 15 Mc/s. Their R103 is displaced by a new design in which good use is made of a 20th Century precision c.r. tube with the high sensitivity of 14 V/cm.

In contrast to these is the Cintel "Synchroscope" without amplification (sensitivity, 60 V/cm) and thereby able to respond to frequencies up to 1,000 Mc/s (−3 dB point).

Although described as an oscilloscope, the Marconi Instruments TF 1104 is much more, since it includes a signal generator covering Bands I, II and III and appropriate v.f. and i.f. ranges, and crystal-calibrated marker pips; in fact, all required for viewing television and v.h.f. frequency characteristics.

To demonstrate the present possibilities of transistors in this field Mullard showed an experimental oscilloscope providing all the usual facilities of the simpler kinds of instrument (see block diagram), with no thermionic valves and a total power input of 6 watts at





Block diagram of Mullard experimental oscilloscope.

12 V. Another interesting transistor unit is a B.T.-H. time-base calibrator. It generates an output super-imposing a time scale on any oscilloscope screen, with small divisions at 10 μs intervals, double-size at 50 μs, and triple at 100 μs, the whole being controlled by a 100-kc/s, $\pm$0.01% crystal.

A delay circuit enables the whole width of the screen to be used for observing any small section down to 10 μs of a waveform up to 2,000 μs in duration and measuring it to within 0.5 μs. Other accessory equipment includes the Mullard television line selector for displaying any individual line on an oscilloscope, the Furzehill two-way beam switch for converting a single-trace oscilloscope to a dual-trace type, and the Cossor monitor with separate c.r.t. and brightness and focus controls for duplicating the trace of an oscilloscope to enable one to be observed while the other is photographed.

A new attenuator was shown by Advance Components: the A64 for a.f. and low r.f., giving a range of 0 to 70 dB in steps of 1 dB at 600Ω. A series of fixed attenuators by Hatfield Instruments, consisting of disc and rod resistors arranged in coaxial "T" formation have the exceptionally wide frequency range of 0-1,000 Mc/s; the standing wave ratio in a 72-Ω line at the highest frequency is given as 1.1, and the inaccuracy of attenuation $\pm$0.3 dB. One might have thought that the now well-known Sullivan decade air capacitor would have satisfied all requirements as regards setting accuracy, but the very small residual errors are now corrected on the 100 pF steps by an auxiliary set of curiously shaped sectors. The same firm showed an example of their new mutual-inductance standards, temperature-compensated in a similar manner to the Sullivan-Griffiths self-inductance standards. Decade inductance boxes are less commonly available than resistance or even capacitance boxes, but Muller-Barbieri (Switzerland) showed one of three models, giving 1 mH to 10 H in 1 mH steps, each step being a toroidal coil on a molybdenum-permalloy powder core.

New bridges were not much in evidence; one of the exceptions was the Avo Universal Measuring Bridge, with 24 ranges of R, C and L, covering almost all component values except inductances below 1 mH. The source is an internal 1-kc/s oscillator for C and L, and d.c. for R; and the indicator is a meter. Capacitor leakage can also be measured; appropriate test voltages are provided. A Pye inductance bridge is interesting for its direct-reading scale, with thumb-operated main and fine controls strongly resembling a well-known type of radio tuning control. The indicator is a miniature c.r.t., which facilitates phase and amplitude balance. Range 0.1 mho to 10 mhos with accuracy at worst $\pm$0.2% of reading. A deviation test bridge shown by B and K has provision for interchangeable meter scales marked with appropriate limits. Not only can production components be rapidly tested against standards, but also phase angle against a reference voltage. The advantages of transistors for bridge oscillators and balance indicators were shown in a number of exhibits by Tinsley.

Not only for ordinary oscillators but also (in conjunction with quartz crystals) as frequency standards are transistors being applied. A good example is the Labgear oscillator, giving a single standard frequency up to 10 Mc/s, with point-contact transistor, complete with battery in a case $3\frac{1}{4}$in $\times$ 2in $\times$ $1\frac{1}{4}$in overall.

A somewhat similar "potted" calibrator by Elliott is stable to better than 1.5 in 10^6 per °C. With a 100 kc/s crystal, useful harmonics are available up to 30 Mc/s. For low frequency standards, Elliott have developed transistor-maintained tuning forks. The Furzehill G410A frequency standard uses a 5-Mc/s crystal with valve multivibrator dividers to provide pulses at 5 Mc/s, 1 Mc/s, 100 kc/s, 10 kc/s, 1 kc/s and 100 c/s, modulated if desired at 400 c/s, and selected by push buttons. A Labgear prototype standard contains an oven-stabilized 200-kc/s crystal and a t.r.f. radio receiver tuned to the B.B.C. 200-kc/s carrier, enabling the source of frequency to be known within about 1 in 10^8. Finally, a version of the famous G.P.O. frequency standard with a daily stability of better than 5 in 10^{10} was to be seen on the Airmec stand.

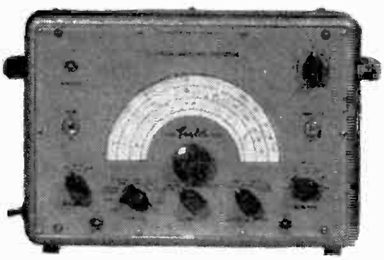
Coming now to signal generators and considering them in ascending order of frequency we have first the Dawe 442 v.l.f. oscillator, 0.1 c/s to 10 kc/s, notable for the placing of the main heat-producing components —the valves—more or less in the open air, though protected by recessing. Another interesting feature is the provision of two outputs in quadrature. The Wayne-Kerr wide-range a.f. oscillator now appears at a further stage of development, the most interesting feature being its frequency and amplitude controls with scales arranged for fool-proof direct reading; the frequency control combines the advantage of continuous variation and main-point working. The instrument meets Ministry requirements with a clearly visible mains-voltage adjustment mounted on the front panel under a transparent slip. The Salford 50 c/s to 50 kc/s signal generator is unusual in being provided with crystal calibration, checking by which is facilitated by scale markings. For use with their Acoustic Calibrator, Dawe have produced a compact transistor oscillator to provide a signal at 400 and 1,000 c/s, but it can also be used as a bridge source, etc.

The Muirhead D783 Square-Wave Shaper, for use in conjunction with sine-wave sources in the range 20 c/s to 300 kc/s (or with 50 c/s from its own power supply for that frequency) gives an output with a rise time less than 0.08 μs. This parameter is of even greater importance in pulse generators, of which the Nagard Type 5001 is a new example. The recurrence frequency of the internal oscillator can be varied from 0.1 c/s to 1 Mc/s, and pulse width and delay from pre-pulse are variable from 0.2 μs to 2s.

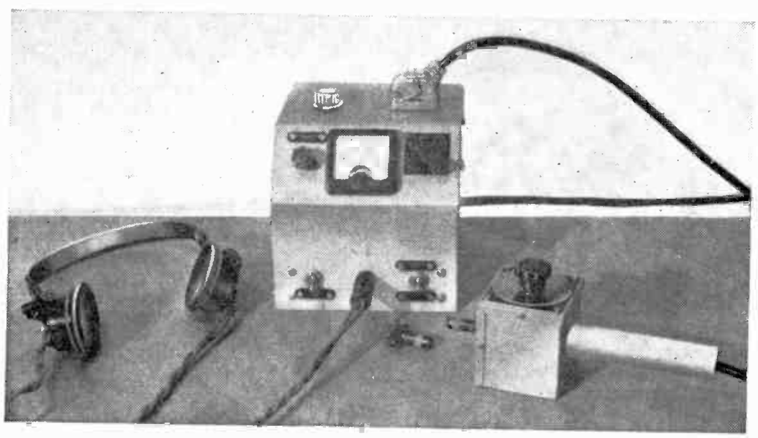
In new r.f. signal generators the accent is on v.h.f. The Avo AM/FM model already mentioned is a modified production form of the prototype seen last year, and covers 5 to 220 Mc/s in eight ranges with a.m. and 65 to 120 Mc/s with f.m. Special arrangements are included for increasing the accuracy of the scale close to any desired frequency. The Airmec Type 204, covering 2 to 320 Mc/s, is notable for having nine different modulation facilities, including a.m. or f.m. or both together, internal or external, or pulse modulation. Two v.h.f. signal generators were shown by Hatfield: the LE120C covering 3 to 300 Mc/s; and the LE250B for 3 to 20 Mc/s and 81 to 105 Mc/s intended for production testing of Band II f.m. receivers. Modulation is by either sine or sawtooth waveform, the latter being intended for display of selectivity and discriminator curves. A special feature of the Marconi TF1066 FM/AM signal generator is an incremental tuning system in which small frequencies are read from a meter with a direct calibration valid at all carrier frequencies.

The lower microwave frequencies are covered by another Marconi signal generator, the TF1058, built around a klystron oscillator of a new type in which the length of a coaxial line and the reflector voltage are simultaneously varied by the frequency control to enable the unusually wide range of 1.7 to 4 kMc/s to be covered in one band. A similar technique is adopted in the Sanders XT312 and XT314 signal generators which between them cover 1.5 to 12 kMc/s. Again, the direct-reading wide-band frequency control is of particular interest; linearity and accuracy of the scale is achieved by an adjustable cam controlling the coaxial piston and geared to a reflector voltage control. A wide range of microwave equipment shown by Sivers Lab (Sweden) included a signal generator which, like all their variable-frequency equipment, was provided with a cyclometer type of frequency indicator direct reading in Mc/s. For measuring noise factor of X-band receivers, the Marconi TF1070 noise source consists of a discharge tube mounted in the E-plane of a length of wave guide, generating a standard noise signal 15.5 dB above thermal noise.

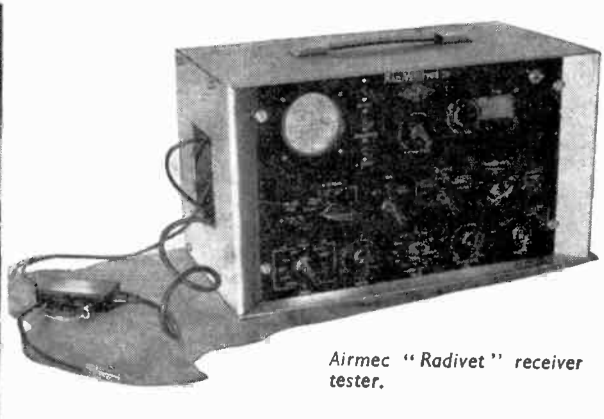
Other microwave gear by Sanders included a No. 16 waveguide test bench (8.2 to 12.5 kMc/s), a calibration receiver with first detector crystal mounted on a three-stage low-noise head amplifier, and a high-power dummy load. B.T.-H. equipment included an X-band automatic s.w.r. and impedance measuring equipment in which a Smith chart display is given on a c.r.t. screen, an X-band dielectric test set for relative permittivity and *tan* δ, and a Q-band "Enthrakometer" for measuring power at 35 kMc/s by the change in resistance with temperature of a very thin platinum film suspended across the incoming waveguide. Another solution of the power-measuring problem, applicable to all frequency bands, is the E.M.I. Precision Microwave Wattmeter Type 1, which operates in a novel manner, making use of what is in effect an a.f. phase-shift oscillator amplitude-stabilized in the usual way by a thermistor. In this case the thermistor is used as the r.f. termination, and the rise in temperature due to the power being measured is automatically compensated by a change in a.f. current through it; this current is therefore a measure of the r.f. power dissipated in the thermistor, which is
(*Continued on page* 279)



Taylor Type 94A television alignment generator.



Right:—Cintel "Signal Tracer."



Airmec "Radivet" receiver tester.

Left:—Microwave equipment made by Sivers Lab.

kept at practically constant resistance. Favourable accuracy and ease of control are claimed. The power range is 0-4 mW in steps of 0.1 mW. A thermistor is also the basis of a method by Microwave Instruments in which it forms one arm of a Wheatstone bridge, calibrated by d.c. substitution.

The highest-frequency microwave equipment seen was an experimental test bench for 75 kMc/s (4 mm wavelength). When it is considered that at this frequency a half-wave dipole aerial would be only 2 mm long some of the instrumental difficulties can be imagined! One of them is the production of such waves; actually an 8 mm klystron was used, with a silicon crystal to generate the second harmonic. Very close tolerances, of the order of 0.0001in, are necessary in the "plumbing." The usual slotted-line type of s.w.r. indicator being impracticable, a rotary indicator is used, the phase of the standing wave being measured directly as a physical angle.

Another fascinating microwave exhibit was the Polarad spectrum analyser, which was shown displaying the spectrum of a pulse-modulated signal in the 0.91 to 4.56 kMc/s band. This consists of a series of vertical lines (carrier and sidebands) spaced at frequency intervals equal to the pulse recurrence frequency and having an envelope width dependent on the pulse width.

Lastly, there were exhibits not falling clearly into any of the foregoing categories; for example, a selection of the Heath (U.S.A.) instrument kits described by C. B. Bovill in the October 1955 issue of *Wireless World*. An interesting feature of the valve-voltmeter kit is the use of an etched circuit. The Taylor 94A Television Waveform and Alignment Generator provides an exceptionally comprehensive selection of signals for testing television, f.m. and short-wave receivers, including a number of television patterns—Continental and American as well as British standards— with synchronizing, interlacing, etc., signals, c.w. with and without a.m. or f.m. at various output levels, a wobbulated signal for alignment in conjunction with an oscilloscope, and a variable-output a.f. signal. The r.f. frequency coverage is 8 to 230 Mc/s. As a complement to the well-known Airmec "Televet" there now appears the receiver tester Type 211 or "Radivet," giving a r.f. signal in the frequency ranges 0.1 to 15 Mc/s and 85 to 100 Mc/s with a.m. or f.m. or both, and an a.f. signal 100 c/s to 15 kc/s. A crystal calibrator is incorporated for accurate frequency checking, and an oscilloscope for wobbulator display or general purposes. A considerable amount of signal and circuit testing can be performed by the Cintel Signal Tracer, consisting of a probe unit containing a calibrated oscillator covering 0.1 to 100 Mc/s with plug-in coils, and a power unit containing the indicating meter, working on the "grid-dip" principle. For recording audio frequency characteristics, B & K showed equipment consisting of a manual or motor driven beat-frequency oscillator in conjunction with an output recorder with a selection of 10 paper chart speeds from 0.003 to 100 mm/s.

Production test equipment included a shorted-turn detector by Nash and Thompson capable of indicating one shorted turn in a coil of 40 s.w.g. wound to a radial depth of ¾in; continuity of the coil is also shown. The object is to detect faults in transformer coils before assembling the cores. For automatically inspecting strength of permanent magnets at high speed, B.S.A. showed test gear working on the principle of upsetting a balance by saturation of one of two small mumetal rods wound with coils forming twin arms of a bridge. A relay operates a rejection mechanism for magnets below standard. A B.T.-H. production tester for point-contact germanium diodes displays on a c.r.o. the forward and reverse current/voltage characteristics, simultaneously but with appropriately different scales; and by applying a square wave to the diode under test enables recovery time (hole-storage effect) also to be assured. The same firm exhibited a transistorized direct-reading frequency meter of the charge-discharge type covering 0.3 to 100 kc/s f.s.r., for signals of any waveform and any amplitude between 0.1 and 500V.

A rather unusual instrument, designed for use in a.c. testing of magnetic materials is the form-factor meter by Tinsley, having two ranges: 1 to 1.5 and 1 to 3. The Airmec phase-meter Type 206 enables both phase and gain to be measured in the frequency range 20 c/s to 100 kc/s. For amplitude modulation in signal generators and similar instruments, the Hatfield balanced modulator Type LE90A is claimed to be completely free from frequency modulation. It makes use of the firm's wide-band r.f. transformers in conjunction with a balanced pair of germanium diodes, and covers frequency ranges of 0 to over 100 Mc/s for the modulating signal and 3 to over 400 Mc/s for the carrier. Finally, one of the largest instruments, the Ellott "A-PAP" microwave aerial near-field phase and amplitude plotter. In contrast to radiation-field polar-diagram equipments, which require a fair amount of space and precautions against reflections, and which do not necessarily indicate the cause of any undesired feature in the radiation pattern, this measures the field close up across the front of the aerial being tested. It consists of a r.f. source to feed the aerial, and a pick-up automatically traversed, the output amplitude and phase of which are recorded on a chart.

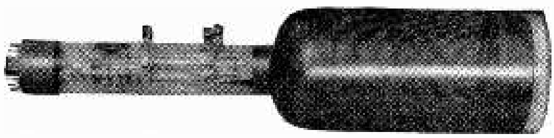
NEW VALVES AND SEMI-CONDUCTORS

"Glassware" and Allied Exhibits at the R.E.C.M.F. and Physical Society's Shows

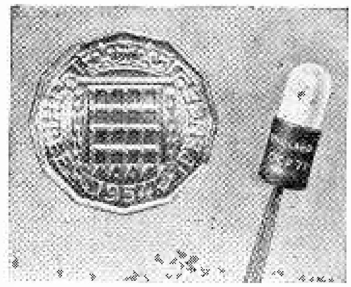
Special and Transmitting Types.—The use of ceramic instead of glass for valve envelopes is not new, but appears now to be on the verge of greater development. It gives greater mechanical strength, smaller size for a given power dissipation, enables the valves to work at higher ambient temperatures and permits more effective de-gassing during manufacture so that greater emission current can be obtained under pulse conditions. A series of ceramic valves shown by Ferranti included triodes for oscillators and amplifiers, with anode dissipations ranging from 15 to 100 watts; low-μ triodes for use in stabilized power packs, and indirectly heated half-wave rectifiers. The maximum operating temperatures are in the region 180°-250°C.

A somewhat larger ceramic valve was shown by English Electric. This was a coaxial transmitting tetrode, CR1101, with an air-cooled anode capable of dissipating 2 kW. The valve will take full ratings up to as high as 900Mc/s and at this frequency will give a useful c.w. power output of 600 watts. English Electric were also showing a new magnetron, M541, designed for pulse operation at about 1,200 Mc/s. It is mechanically tunable over a 10% frequency band and will give a pulse power output of 0.5 megawatt.

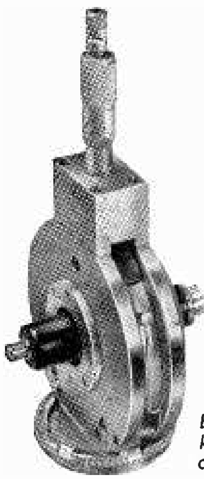
The backward-wave oscillator is one of the latest types of velocity-modulation valves and is comparable with the ordinary travelling wave tube except that the r.f. field energy travels in the opposite direction to the electron beam. A notable feature is the wide frequency variation which can be obtained by altering the beam accelerating voltage, and in the Mullard valve on show, type MS1203, this amounts to 7,000 Mc/s-11,500 Mc/s (with 50 mW output). A more familiar velocity-modulation oscillator valve is the klystron, and this is not normally known as a tunable device. However, E.M.I. Electronics were showing a new reflex klystron,

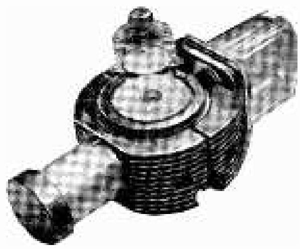


20th Century Electronics tube with spiral p.d.a. electrode.



Above :- Mullard photo-transistor compared in size with a threepenny bit.

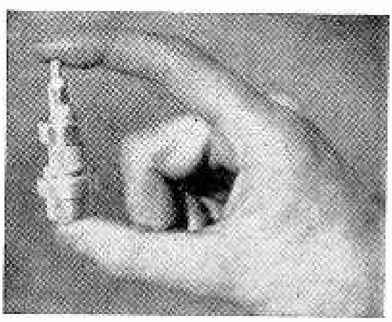




English Electric tunable magnetron.

Left :- G.E.C. r.f. heating power valve.

E.M.I. Electronics reflex klystron in tunable cavity.



Ferranti 15-W ceramic triode for use up to 2,000 Mc/s.

R5222, which operates with an external cavity resonator tunable between 8,400 Mc/s and 10,300 Mc/s. The power output rises from 28 mW to 55 mW at mid-band frequency and falls again to 25 mW.

Receiving Types.—In multichannel television reception the receiver often has to handle signals of widely different amplitudes, and one unfortunate result of this can be cross-modulation distortion between vision and sound, giving an objectionable buzz on sound. G.E.C. have now produced a new variable-mu pentode (Type W729) for television i.f. amplifiers, which minimises this distortion while at the same time maintaining a high slope and giving a better control characteristic than conventional variable-mu valves. For a.m./f.m. reception this firm also displayed a new triple-diode triode DH719/EABC80 with one diode having a separate cathode, while Mullard had a variable-mu r.f. pentode with two diodes (EBF89), which is notable for a low anode-to-grid capacitance (less than 0.002 pF) and reasonably high slope (3.6 mA/V); this makes it possible to obtain high gains without instability in the i.f. stages of a.m. or f.m. receivers. Another new Mullard valve for a.m./f.m. reception was the miniature battery r.f. pentode DF97, intended for portable sets.

Transistors.—A new kind of junction transistor which overcomes the inherently slow response of the conventional junction type is the "avalanche" transistor, and an experimental model was shown by Mullard. The action depends on the fact that the collector is operated at a higher voltage than normal, so that the holes flowing towards it reach high velocities and collide with germanium atoms nearby, producing additional holes and electrons. These in turn produce more current carriers by the same effect, and a sudden multiplication, or "avalanche," occurs which results in a very rapid build-up of current through the transistor. Moreover the high collector voltage causes the hole-depletion layer to stretch almost all the way through the base to the emitter and consequently the holes only have to diffuse a very short distance through the base before being swept into the collector. The avalanche process is analogous to the breakdown in a cold-cathode discharge valve except that the base retains control during the conduction. The device is particularly suitable for pulse applications, for it will handle rise times as short as 0.01 μsec.

A power transistor (germanium junction) that can be used to switch as much as 5 amps was shown by B.T.-H., while G.E.C. had a new junction type, EW53, in which the base is connected to the metal can to give lower thermal resistance and improved mechanical strength. English Electric have now entered the field with some junction transistors for low-power audio applications, while Brimar have introduced three new types for similar applications, TS1, TS2 and TS3, which are hermetically sealed and replace their earlier TJ types.

Rectifiers.—The most important development in this field has been the emergence of the silicon junction diode. This device is characterized by its small size for a given current rating, its ability to work at high temperatures and its low reverse current. A typical small unit about the size of a $\frac{1}{4}$-watt resistor might give a rectified current of 200 mA and have a reverse current of 5 μA with a peak inverse voltage of 100V while operating at a temperature of 100°C. Examples were shown by B.T.-H., Ferranti, G.E.C., and S.T.C. Heavy-duty types capable of passing currents of 30-50 amps at 100-200 volts were also on view and these were still only about $\frac{3}{4}$in in diameter.

Germanium junction rectifiers, by comparison, have the advantage of lower forward resistance. As an example of what can be done with them, G.E.C. were showing a water-cooled germanium diode capable of giving a rectified current of 300 amperes! Mullard had a new germanium junction diode, Type OA10, notable for its extremely low hole-storage characteristics.

A new method of depositing selenium on metal plates by a vacuum evaporation process has led to metal rectifiers of somewhat improved performance. The process gives a more intimate bond and makes possible a thinner layer of selenium, as a result of which the forward resistance is lowered and the current rating for a given size of rectifier is increased by about 25%. The life of the rectifier is also said to be improved.

A range of selenium rectifiers notable for their ability to work at ambient temperatures as high as 85°C was displayed on the Westinghouse stand. One type suitable for television receivers, giving 300mA at 270V, has elements contact-cooled at the edges on to the aluminium case and is "potted" in a block of resin of about 1 cubic inch. It will operate successfully even when the chassis to which it is fixed is as warm as 60°C.

Photo-electric Devices.—An improved version of the conventional "venetian-blind" photo-multiplier tube was shown by 20th Century Electronics. In this a carefully designed electrode system gives some degree of focusing from stage to stage, and so leads to improved extraction of slow secondary electrons and higher inter-stage efficiency. Increased efficiency was also the feature of the new cadmium sulphide photocells exhibited by B.T.-H., which are several orders of magnitude more sensitive than selenium cells. Both single-crystal and powder-layer types were shown.

Another new material which has been found to have photo-electric properties is germanium, when used in junction devices, and an embodiment of this was the photo-transistor, type OCP71, shown by Mullard. It is very small and will operate from a low voltage (about 10 V), giving an output current when illuminated in the region of 5 mA. The "dark" current is not more than 300 μA.

C.R. Tubes.—For television the 21-inch tube with a 90° deflection angle is the latest thing on the market and two new types were on show—the Brimar C21KM, with a tetrode gun, and the G.E.C. 7501A with a triode gun. Both have aluminized screens. As a contrast, Ediswan have reverted to the 9-in screen size in their new tube CRM93—although they have also produced a 24-inch rectangular type.

Among oscilloscope tubes 20th Century Electronics were showing a new type which has the interesting feature of a spiral post-deflection acceleration electrode instead of the usual series of rings. It is formed by a coating of resistive material on the glass and the accelerating potential (10 kV) is applied across it. This gives a potential gradient which increases evenly towards the screen and so avoids the lens effect which occurs between separate rings and also the need for a series of separate acceleration voltages. The writing speed of the tube is 1,000cm per microsecond. Another tube with post-deflection acceleration, the Cintel G601-C4, is capable of recording frequencies as high as 1,500 Mc/s.

LETTERS TO THE EDITOR

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

"*Precision Photographic Timer*"

MAY I reply to your correspondents, Messrs. Hercock, Neale and Askew (April, 1956)? Changes in contact potential are usually a limiting factor in the design of low-frequency apparatus such as the photo timer. The input voltage e_{in} was chosen to be a compromise between high values giving little or no increase in CR and low values where contact potential changes are relatively more significant. In regular use, the grid voltage of a valve such as the EF37 will return to within 100 mV of its mean value and, with intermittent use, to within 200 mV, falling to 100 mV in less than an hour. With the suggested value of 4.7 V for e_{in} 100 mV corresponds to only about 2% change in interval. Larger changes due to ageing may be allowed for using the 2.5-K potentiometer (Fig. 4).

Changes in the integrator valve internal gain have only a small effect on the interval (conventional negative feedback theory). The effect becomes more important as e_{in} is reduced, due to the relatively greater significance of the changed grid voltage excursion, but this is a price paid for the effective increase in CR, as with the contact potential changes.

The steps of $\sqrt{2}$ (or 1.58) are logarithmic, which seems more appropriate than arithmetic steps of, say, 15 seconds, and though admittedly rather coarse, were chosen for two reasons—(*a*) a large total range had to be covered using a standard 11-way switch, (*b*) most modern printing papers (such as Ilford's Plastika) are supplied with a leaflet advising that there is no harm in using developing times of −50% to +100% of the optimum if desired—this is more than adequate to interpolate between steps of 1.58 in exposure time.

Measurements on an Atmite disc type L275702A show a temperature coefficient of 0.23 μA per °C measured at 27 V, 40 μA (cold). Thus a 1°C rise will increase e_{in} by 0.58%. My own timer is constructed on a conventional tinplate chassis, completely enclosed in a plywood box. The temperature difference of the chassis underside rises on a 50 min. exponential (approx.) to an asymptotic value of 16°C (approx.). If used immediately after switching on, the timer would therefore give intervals 9.2% longer than in the temperature-stabilized condition. On this account it appears advisable to allow about 30 min. warming-up time where it is desired to keep to better than 5% accuracy.

G. A. Askew is quite correct in stating that the Miller integrator valve can stay on its grid base without an anode load (for negative "run-down" only). In the design of a single valve Miller time-base to work faster than about 20 V/μsec the feedback capacitor current and the usual load capacitor current are, of course, larger than the anode resistor current. I should have mentioned that the only reason the 1.5-M anode resistor is used in my circuit is to set up the recommended conditions for low grid current (cf Mullard ME1400). The recommended reduction of the heater voltage to 4.5 V has been noticed by myself and others to be ineffective and possibly harmful to valve life.

J. G. THOMASON.

Malvern, Worcs.

Single or Double Sideband?

BEFORE the opening of the new London television station I wrote to the B.B.C. in terms similar to those used by your correspondent, "Lambda" (April issue).

Nevertheless, the event has proved both of us wrong, although I imagine the B.B.C. were unaware that this would be so. In spite of "Lambda's" curves, I at least am able to receive a satisfactory picture using a receiver turned to the upper sideband. In fact, it is possible to resolve the 2.5-Mc/s bars on Test Card C (previously I could just see the 3-Mc/s).

I am not suggesting that the B.B.C. were right in their decision to use the vestigial sideband system, and it would obviously not have been possible to consult the viewing public. It is perhaps fortunate for the B.B.C. that the thousands of U.S.B. receivers still in use were not rendered obsolete overnight.

F. R. ESTALL.

South Benfleet, Essex.

B.B.C. Reply

YOUR correspondent, "Lambda," sets out in the April issue some of the arguments against the change to vestigial sideband transmission that has been made with the opening of the Crystal Palace television station. These points were all considered, but the factors that principally influenced the decision were:

1. The desirability of using the same type of transmission at all B.B.C. television stations; all the permanent post-war stations have a vestigial sideband characteristic.
2. The improved power-conversion efficiency and greater power output that could be achieved in the transmitting equipment.
3. The improved performance of the transmitter and aerial system over the band transmitted.
4. The saving of 2 Mc/s of spectrum space, which may possibly be useful for some other purpose in the future.

Before the decision was made, tests were carried out, with the co-operation of the radio industry, on a number of commercial receivers designed to favour the upper sideband. The tests were made with a filter rather more severe in attenuating the upper sideband than the actual characteristic of the Crystal Palace transmitter. It was found that the loss of detail in the picture was much less than would be expected from "Lambda's" theoretical diagram. Even without any modification of the receivers the resolution was only slightly degraded, and the difference was barely noticeable on most of the receivers tested.

The number of receivers favouring the upper sideband that are still in use is certainly less than the figure of 60,000 quoted by "Lambda" from Sir Noel Ashbridge's paper of 1951 as the total number of sets in use at the time to which he referred (shortly after the reopening of the Alexandra Palace station after the war). Nearly all the receivers of this type are at least six years old, and many considerably older; a number of them have, in any case, been replaced by sets capable of receiving the I.T.A. transmissions in Band III, which, incidentally, also have a vestigial sideband characteristic.

The B.B.C. regrets that the owners of some of the older receivers favouring the upper sideband may need to modify them, but believes that such cases are few and that the decision was justified in the general interest.

E. L. E. PAWLEY.

Head of Engineering Services Group, B.B.C.

Helping the Blind

PRIOR to losing my sight in 1946, I was a regular reader of *Wireless World*. For the past eight years I have been employed by Decca as a mechanical inspector, using instruments specially adapted for the use of the blind. I have now become interested in electronics, and particularly in tape recording, which I use mainly for correspondence with other recordists in all parts of the world. I am a member of two tape clubs, both centred in the U.S.A. I would be interested to hear of any similar British club.

Many of my correspondents are blind, and you will appreciate that our greatest difficulty is to keep abreast of developments and techniques, for there are no electronics or radio technical journals published in Braille in this country; in fact, the only one published anywhere in the world is the *Braille Technical Press* in New York.

You will appreciate that when an article is read aloud by someone not interested in the particular subject, the resultant impression can be confusing and often misleading.

In the U.S.A. this has been overcome to a large extent by club members reading articles, together with their comments, on to tape and sending them to fellow members for discussion.

May I suggest that some of your readers might like to participate in such a service over here, and regularly read items and articles from *Wireless World* and so help us to keep up to date, and to overcome to some extent our handicap.

Also, may I ask for suggestions for an aural modulation indicator to substitute for the usual visual indicator on tape recorders?

CHARLES H. STANDEN.

London, S.W.9.

Terminology; Warlike—

I WISH to make a complaint. Why must we be *fired at* and *triggered-off* in television circuits? Are we required to assume that the authors have all had Army, Navy or Air Force training?

Cannot we leave these matters to the war-lords and simply speak of the *electro-activation pulse*.

A. HARDWICKE.

Glossop, Derbyshire.

—and Misty

IT seems to me that the terms "demodulate," "demodulator," are bad ones and ought to be dropped, as their meaning is the reverse of that intended. If the windscreen of my car becomes *misted* I switch on a *demister* which removes the mist and leaves the windscreen clear. Yet some radio people, when they have a carrier which is *modulated*, use a *demodulator* to remove the carrier and leave the modulation. This is surely as absurd as saying that a demister is a device for removing the windscreen and leaving the mist.

P. E. K. DONALDSON.

Cambridge.

Tape Amplifier Design

I HAVE recently constructed tape recording equipment based on the design of A. F. Fischmann, published in *Wireless World* as long ago as November, 1954.

It is difficult to understand the use of an 8-μF electrolytic capacitor, C_{55}, to decouple the screen of V_6. Unless the polarizing current is much below average it will be sufficient to reduce seriously the screen voltage, and a capacitance of 1 μF would be adequate.

The record amplifier has a very high degree of feedback, and two identical intervalve couplings. An amplifier to Mr. Fischmann's specification showed a peak of +12 dB above the mid-band gain at 0.2 c/s. This leads to a tendency to instability. The peak can be reduced to +2 dB by the use of the following component values:—C_5, 0.005 μF; C_7, omitted; C_8, 0.25 μF; R_8, 1 MΩ.

The high-frequency peak is smaller, due to the reduction in feedback by C_1, and can be dealt with by 100 pF in parallel with R_7.

It was found impossible to replace V_2 by a pentode, as suggested in the text, because of the difficulty of providing a screen supply that introduced sufficiently little phase shift to avoid low-frequency oscillation.

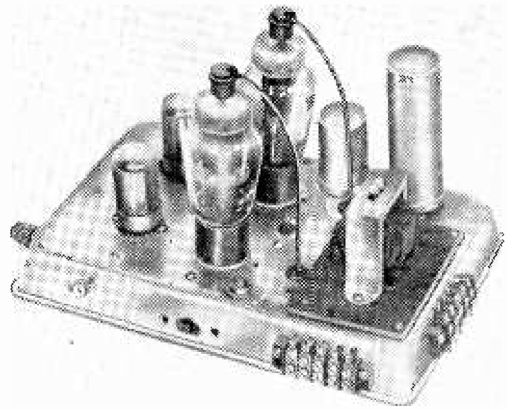
R. C. MARSHALL.

Wheathampstead, Herts.

Cascode A.F. Amplifier

"Long-tailed Cascode Pair" as Combined Pre-amplifier and Phase Splitter

By L. B. HEDGE, Ph.D.



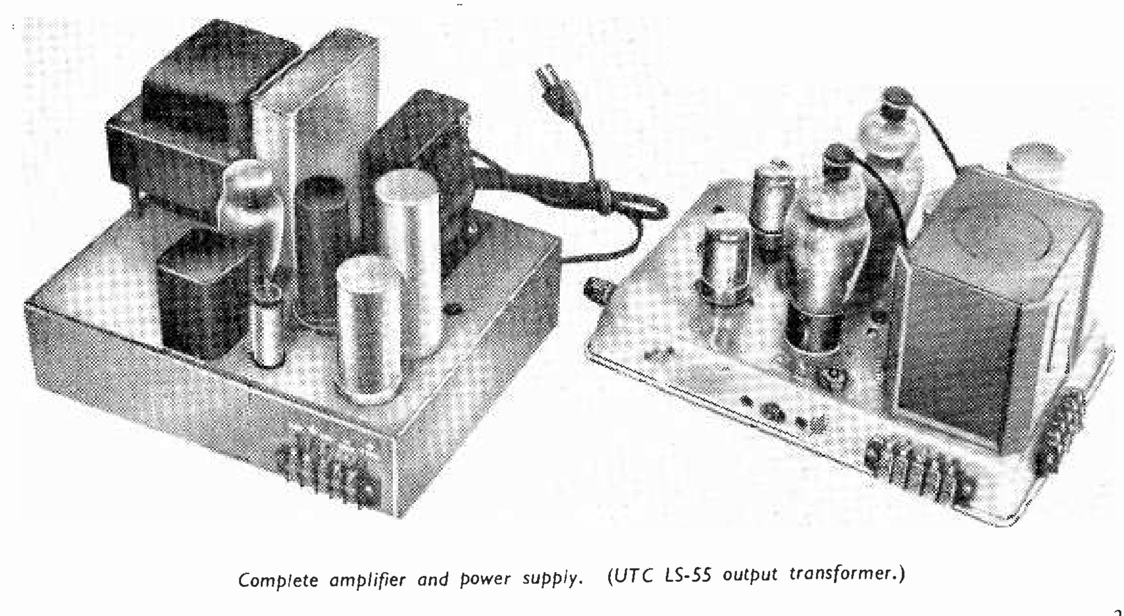
Amplifier with "replacement" output transformer.

THE "cascode" amplifier—a series connection of two triodes which operates much like a single triode, with characteristics practically unattainable in a single triode—has been extensively employed as a high-frequency amplifier during recent years, and more recently as a first-stage, low-level, audio-frequency amplifier (so-called "pre-amplifier"). Although the cascode was developed as a direct-current amplifier for voltage regulator control application[1], its recent uses have been largely based on the inherently low level of stage noise[2]. The importance of minimizing the signal-to-noise ratio in a variety of high-frequency applications, including radar, television and many others, has served to keep attention focused on this low-noise feature as the distinguishing characteristic of the cascode, and its use in the audio-frequency field has also been based largely on this feature.

The amplifier here described (on which patents are pending) is the result of a return to an earlier view of the cascode stage; it is used here because of the characteristics for which it was originally developed—its triode-like performance and its high equivalent amplification factor. Although low noise is no disadvantage in any amplifier, it is of importance only in a stage (the first, barring exceptional circuitry) where the input signal is of sufficiently low intensity to make the signal-to-stage-generated-noise ratio critically small. In the next-to-final stage of an audio-frequency power amplifier, only exceptionally bad design could make the noise generated in the stage a factor of significance in the performance of the system.

High quality in audio-frequency power amplifier performance—uniformity of response and low distortion over the spectrum of audible frequencies—depends in large measure on a few closely inter-related design elements; the output transformer, the feedback circuitry, and the frequency, phase-shift, and attenuation characteristics of the inter-stage couplings which establish the limits within which feedback may be used as an overall corrective[3]. In general the output transformer is the effective limiting element in amplifier performance, and recent impressive improvements have been based on special transformer designs[4].

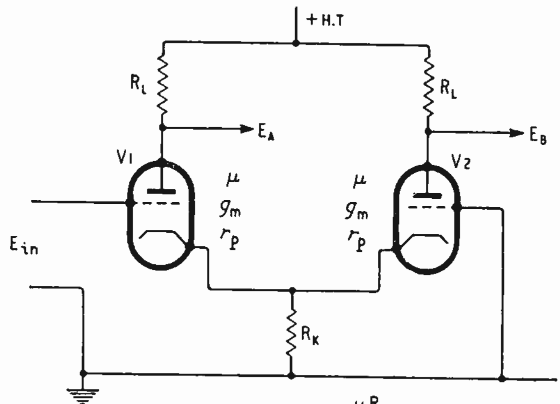
In exploring the problem of evolving an amplifier



Complete amplifier and power supply. (UTC LS-55 output transformer.)

design which would make most effective use of an output transformer of non-critical design—one which would make the best use of any output transformer built into it—it soon became clear that some major changes in " conventional " circuitry would be required. A feedback loop to support a high level of corrective feedback which would include the output transformer and go back at least to the phase-inverter stage seemed a minimum reasonable requirement, and with conventional circuitry this leads to something very much like the basic " Williamson " layout. With the low gain of most popular phase-inverter stages, and the high drive requirements of the output stage, at least one driver stage is required between the phase inverter and the output stage, and an additional stage which may be either before or after the inverter. One direct coupling between stages (as in the Williamson scheme) is quite practicable, but more than one adds serious complications to the power supply and isolation filter problems. The result is a feedback loop which contains two R-C coupling networks and the output transformer, with a possible maximum phase shift of 270°. Stability of the amplifier requires that the loop gain be reduced to less than 1 before the phase shift reaches 180°, and, in view of the phase-shift and attenuation characteristics of the couplings and the transformer, the frequency range over which feedback can be kept high must be considerably smaller than the usable range of the transformer itself[5]. The search for a reasonable way out of this vicious circle of conflicting constraints led to the analysis of the cascode and the cathode-coupled phase-inverter, and finally to the combination of the two—the " long-tailed cascode pair " (l.t.c.p.).

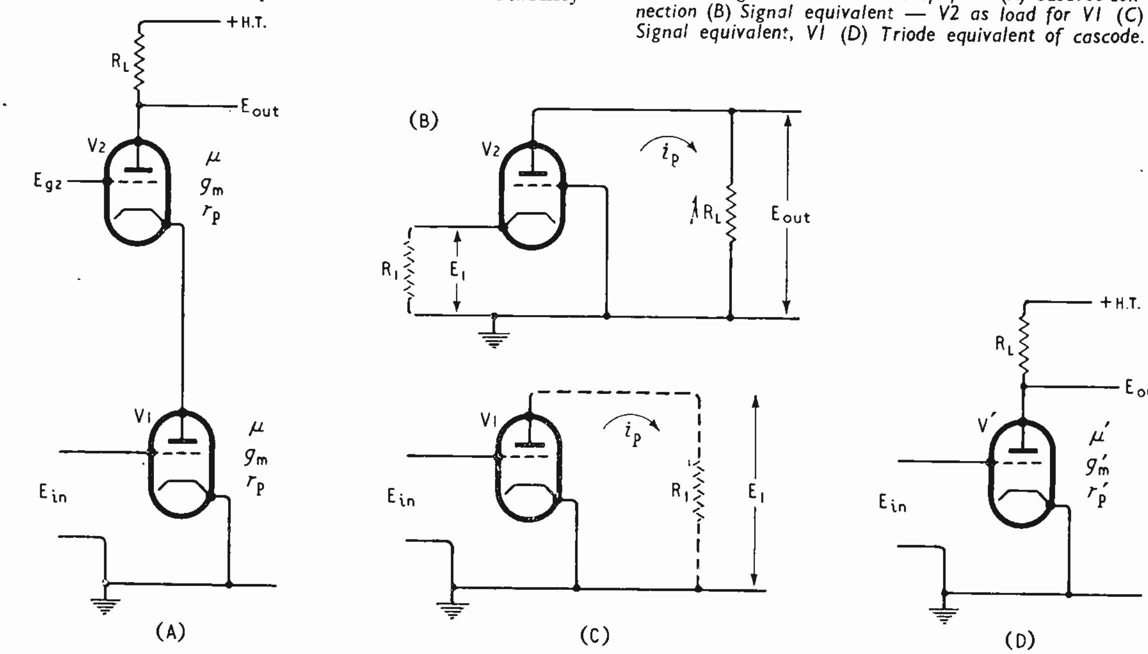
The cathode-coupled phase-inverter is well known and has been extensively used (Fig. 1). The un-by-passed common-cathode resistor provides degenerative feedback to the input tube as well as driving potential for the grounded-grid inverter. The anode-to-anode output of this stage is independent of the value of the



$$E_{AB} = E_{in}\frac{\mu R_L}{R_L + r_p}$$

Fig. 1. Cathode-coupled phase-inverter (long-tailed pair).

Below:— Fig. 2. The cascode amplifier. (A) Cascode connection (B) Signal equivalent — V2 as load for V1 (C) Signal equivalent, V1 (D) Triode equivalent of cascode.



(B)

$$E_1\mu = i_p(r_p + R_L) + E_1 \quad (1)$$

$$i_p = \frac{E_1(\mu+1)}{r_p + R_L} \quad (2)$$

$$R_1 = E_1/i_p = \frac{r_p + R_L}{\mu + 1} \quad (3)$$

(A)

$$-E_{in}\mu = i_p(r_p + R_1) \quad (4)$$

$$i_p = \frac{-E_{in}\mu}{r_p + R_1} \quad (5)$$

$$i_p = \frac{-E_{in}\mu}{\frac{R_L + r_p}{\mu+1} + r_p} = \frac{-\mu(\mu+1)E_{in}}{R_L + (\mu+2)r_p} \quad (6)(3\,\&\,5)$$

$$\frac{E_{out}}{E_{in}} = \frac{i_p R_L}{E_{in}} = \frac{-\mu(\mu+1)R_L}{R_L + (\mu+2)r_p} = \frac{-\mu' R_L}{R_L + r_p'} \quad (7)$$

cathode resistor if the two valves are matched and the anode load resistors are equal, and the ratio of the two anode-to-earth output voltages is[6]:—

$$\frac{E_A}{E_B} = 1 + \frac{R_L + r_p}{(\mu+1)R_k} \qquad (1)$$

Precise balance can be provided by selection of R_k and R_L for given tube characteristics, but if high gain and reasonable power supply voltage requirements are to be realized, μ must be exceptionally large.

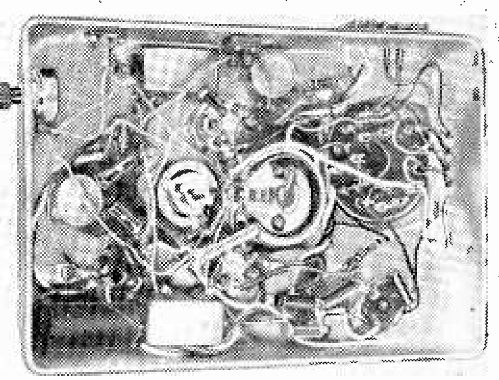
The cascode amplifier consists of a conventional triode with a cathode-driven triode as its anode load (Fig. 2). Analytically the cascode takes the form of a fictitious triode with characteristics μ', r'_p, and g'_m the values of which, expressed in terms of the characteristics of the component triodes (assumed identical) μ, r_p, and g_m, are:

$$\left.\begin{aligned} \mu' &= \mu(\mu + 1) \\ r'_p &= (\mu + 2)r_p \\ g'_m &= \frac{\mu'}{r'_p} = \frac{\mu(\mu + 1)}{(\mu + 2)r_p} = \frac{\mu + 1}{\mu + 2}g_m . \end{aligned}\right\} \qquad (2)$$

Typical twin-triodes in cascode connection should thus provide characteristics as follows:—

Type	μ	r_p	g_m	μ'	r'_p	g'_m
6SN7 7N7	20	7 kΩ	2.9	420	0.15MΩ	2.8
6SL7 7F7	70	44 kΩ	1.6	5000	3.2MΩ	1.6

Anode characteristic curves for these two types were constructed for design reference (Fig. 3). The curves represent measurements on one valve of each type, and may not be good averages in the accepted sense. They do provide, however, an approximate basis for selection of operating points and load-line constructions. Dynamic checks with loads as indicated on the curves and anode supply voltage (E_{bb}) of 475 V

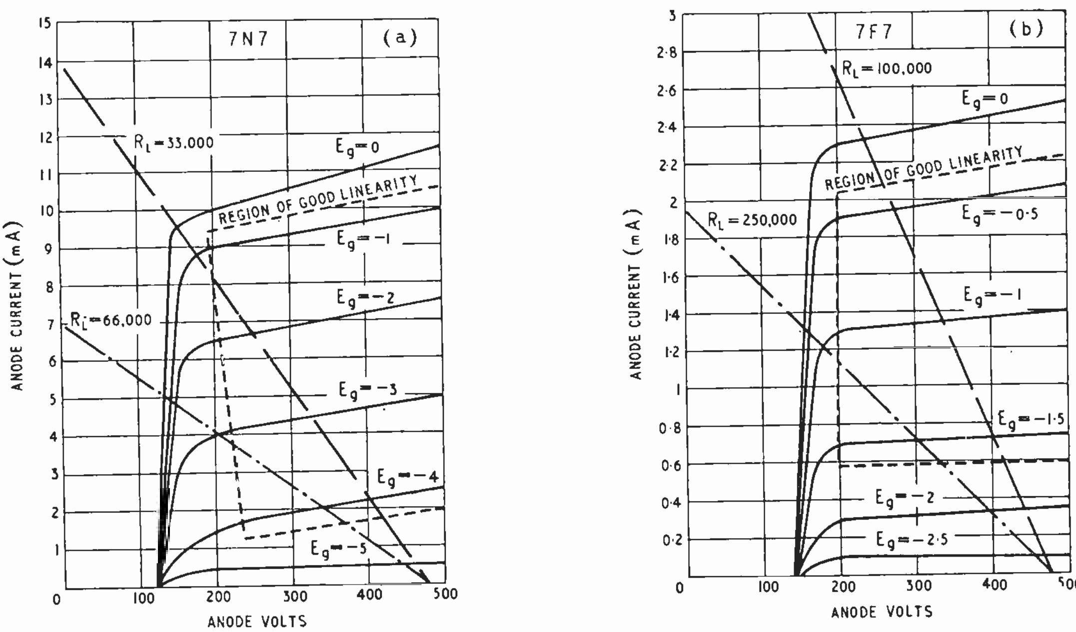


Under-chassis view of the complete l.t.c.p. amplifier.

(the approximate value normally available in an audio-frequency power amplifier, and the maximum available from my regulated adjustable supply unit) check reasonably well with the curves, and even better with the computed values. Within the regions of good linearity to the two cascodes the 6SN7/7N7 should provide a gain of approximately 128 with a load resistance of 66kΩ and a anode supply of 475 volts, while the 6SL7/7F7 should provide a gain of about 360 with a load of 250kΩ and the same anode supply voltage. On the basis of this analysis the experimental amplifier was laid out using 7F7's in the l.t.c.p. stage.

The final circuit of the amplifier is shown in Fig. 4. Type 1625 output valves (12-volt heater versions of the 807—similar in general characteristics to the KT66) were used because they were at hand—as were the

Fig. 3. Cascode amplifier anode characteristics and dynamic check test.



DYNAMIC CHECK – R_L = 33,000, E_{bb} = 475V, E_{g2} = 120V, E_g = −2V, E_{IN} = 0·1V r.m.s, E_{OUT} = 7·5V r.m.s.

DYNAMIC CHECK – R_L = 100,000, E_{bb} = 475V, E_{g2} = 150V, E_g = −1V, E_{IN} = 0·1V r.m.s, E_{OUT} = 14V r.m.s

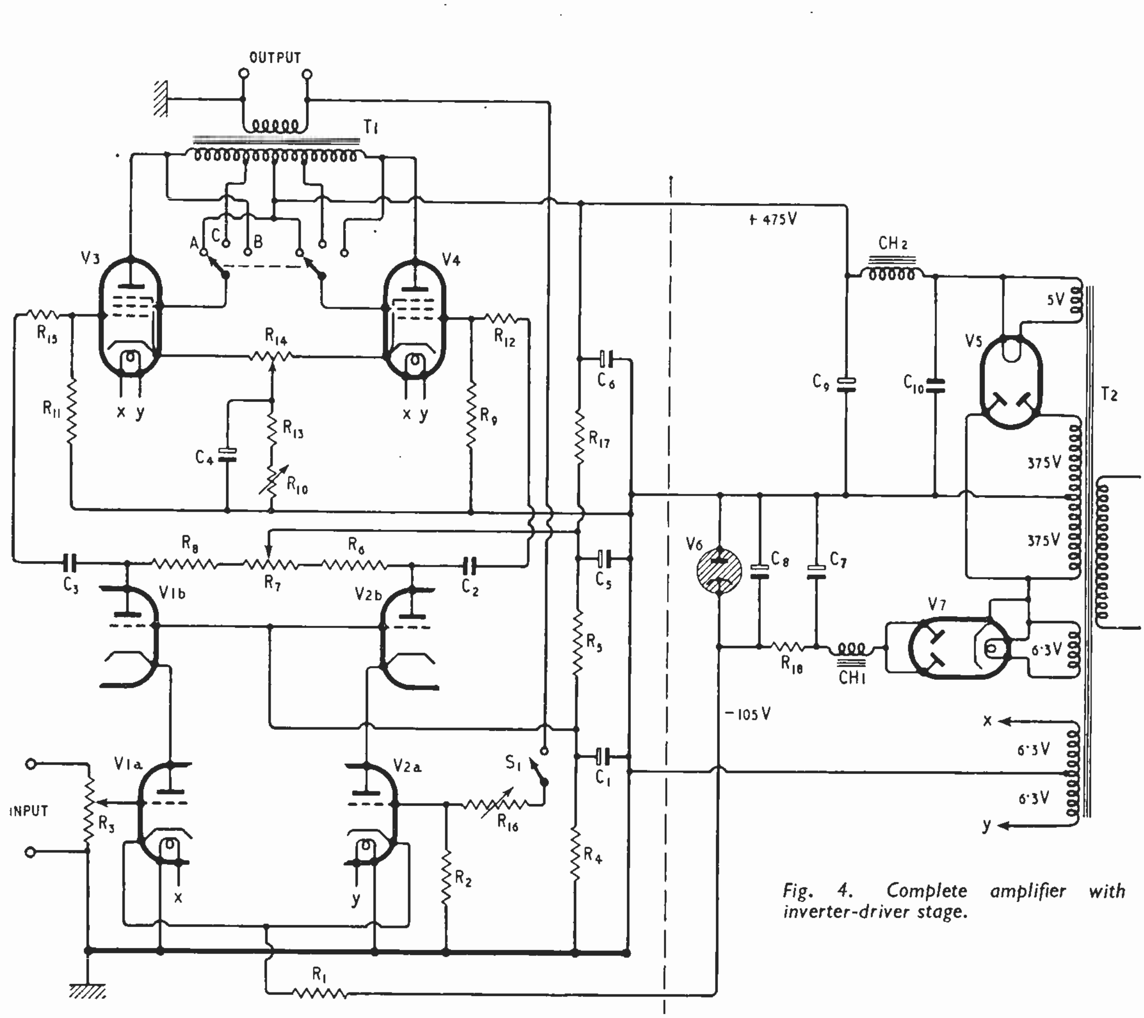


Fig. 4. Complete amplifier with inverter-driver stage.

LIST OF PARTS

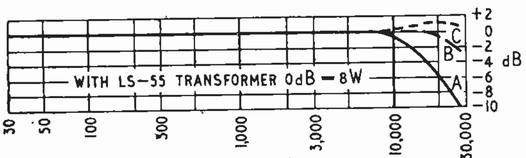
C_1, C_7	20μf, 450V electolytic
C_2, C_3	0.1μf, 600V
C_4	120μf, 150V electolytic
C_5, C_6	40μf, 450V ”
C_8	40μf, 350V ”
C_9	40μF, 450V ”
C_{10}	10μf, 600V
Ch_1	5H, 300 ohm, 40mA choke
Ch_2	10H, 90 ohm, 200mA choke
R_1	50kΩ, 1 watt
R_2	68kΩ, ½ watt
R_3	500kΩ (volume control)
R_4	47kΩ, ½ watt } Cascode 2nd grid voltage divider
R_5	100kΩ, 1 watt }
R_6, R_8	220kΩ, 1 watt } Cascode load and balance adjustment
R_7	50kΩ pot., ¼ watt }
R_9, R_{11}	400kΩ, ¼ watt
R_{10}	200Ω rheostat, 10 watt (Output bias adjustment)
R_{12}, R_{15}	1kΩ, ¼ watt
R_{13}	100Ω, 5 watt
R_{14}	100Ω pot., 5 watt (Output cathode balance adjustment)
R_{16}	1MΩ pot. (Feedback adjustment)
R_{17}	4.7kΩ, ¼ watt
R_{18}	15kΩ, 10 watt
S1	S.P.S.T. switch (Feedback disconnect)
T1	Output transformer—(See text)
T2	Power transformer 375-0-375V, 200mA, heater as required

V1, V2, 7F7; V3, V4, 1625; V5, 574; V6, OB2; V7, 6X4.

7F7's. The essential symmetry of the l.t.c.p. stage suggested immediately the closure of the feedback loop through the grid circuit of the grounded-grid inverter, since satisfactory introduction of the feedback voltage into the input grid circuit is somewhat complicated by the presence of the volume control. Pentode, triode, and so-called “ ultra-linear ” operation of the output stage is provided by the alternative connections (A, B, and C, Fig. 4) for the screen grids of the 1625's.

Performance of the complete amplifier was checked using a United Transformer Company's LS-55 transformer as a reference—a typical “ good ” transformer (reference 7 covers its use in the “ ultra-linear ” connection)—and a “ universal replacement” type, unidentified by manufacturer's name or model designation, culled from the shop “ junk box,” as a kind of “ worst possible ” unit for evaluation of the system. Fig. 5 indicates the effectiveness of the system in providing adequate drive and stable operation at high corrective feedback levels.

The complete amplifier—a “bread-pan layout”[8]—



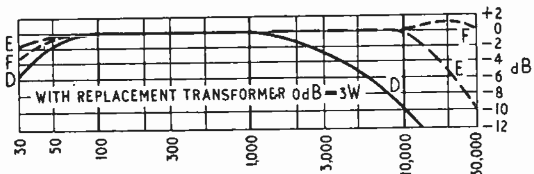


Fig. 5. Performance characteristics of complete amplifier.

Curve A 0dB feedback 0.16V r.m.s. input—U-L connection
0.11V r.m.s. input—pentode connection
Curve B 10dB feedback 0.52V r.m.s. input—U-L connection
Curve C 10dB feedback 0.38V r.m.s. input—pentode connection
Curve D 0dB feedback 0.25V r.m.s. input—triode connection
0.12V r.m.s. input—pentode connection
Curve E 10dB feedback 0.80V r.m.s. input—triode connection
Curve F 10dB feedback 0.40V r.m.s. input—pentode connection

Maximum output watts with harmonic distortion less than 1%:

Transformer Connection	Replacement Triode		Transf. Pentode		LS-55 Transformer U-L		LS-55 Transformer Pentode	
Feedback	0dB	10dB	0dB	10dB	0dB	10dB	0dB	10dB
30 c/s	0.1	0.5	0.1	0.5	12	18	10	15
100 c/s	1	6	1	8	12	18	12	15
1,000 c/s	3	6	3	8	12	18	12	15
10 kc/s	3	6	3	8	12	18	12	15

Note: Increase in feedback from 10 to 20dB with increase in input voltage of approx. x3 changes output characteristics less than 1dB with LS-55 transformer and less than 2dB with the replacement transformer.

is shown in the photographs. As may be surmised, neither construction, layout, nor wiring is critical in any sense. The −105volt supply required for the cathodes of the l.t.c.p. stage is an exceptional requirement, but it is easily met by a simple modification of a conventional power supply, as shown in the wiring diagram of Fig. 4. Since each d.c. connection to the amplifier is to a symmetrical and balanced load, isolation, hum and ripple filter can be quite simple.

The output stage cathode bias scheme shown is simple and effective for providing final stage balance, but it is not in any way a special feature—the Williamson-type network should be equally effective. The cathode bypass condenser in this stage is not necessary either, but the author preters to use it since it tends to reduce distortion if and when the output tubes, by ageing or for other reasons, depart from perfect balance. No provision has been made for static balance of anode currents in this stage, since the author's experience and tests indicate that dynamic balance will produce lower distortion, and that dynamic and static balance frequently occur at different bias adjustment settings.

The "long-tailed cascode pair," by eliminating one inter-stage coupling without reducing gain or seriously complicating the power supply requirements of the conventional power amplifier system, makes the use of output transformers of non-critical design consistent with high quality and exceptional stability. With a real "dog" for an output transformer, this "tail" will wag it so that it will perform like a thoroughbred!

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A.M./F.M. Signal Generators, an e.h.f. spectrum analyser, a v.h.f. alignment oscilloscope, s.h.f. and v.h.f. wavemeters, a counter-type frequency meter, and an f.m. station monitor are among the 25 new instruments described in the Marconi Instruments 1956 "Electronic Measurement" catalogue. From the company's address at Longacres, St. Albans, Herts.

Voltage Stabilizers; a.c. for output currents of 0-9A and 0-30A at mains voltage; and d.c. with outputs of 0-7A, 1-30V and 0-2.5A, 1-15V. These and other electrical control instruments described in a 1956 general catalogue from Servomex Controls, Crowborough Hill, Jarvis Brook, Sussex. Also a data sheet on an i.f. waveform generator suitable for testing servo mechanisms, etc.

Electronic Thermometer, for industrial or medical use, with quick response. Uses germanium thermo-sensitive device and has accuracy of ±1°C. Available in four types: 25°-45°C; −10°-110°C; −50°-160°C; 0°-210°C; manufactured by Ultrakust Geratebau of Germany. Leaflet from the distributors, Headland Engineering Developments, 164-168, Westminster Bridge Road, London, S.E.1.

Decimal H.P. Electric Motors, for sound recording and reproduction equipment. Shaded pole induction motors for 100/130V or 200/250V a.c. Type DHP1: speed 1,345 r.p.m. at 50 c/s; running torque 2 in-oz. Type DHP2D: speed 2,800 r.p.m. at 50 c/s; running torque 3.5 in-oz. Leaflet from The Garrard Engineering and Manufacturing Co., Newcastle Street, Swindon, Wilts.

Silvered Mica Plates for assembling capacitors with relatively simple equipment. Description, diagrams and tables of data in an enginering data sheet from Johnson Matthey and Co., 73-83, Hatton Garden, London, E.C.1. Also similar sheets on contact materials, their choice, properties and availability.

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Waveform Analysers, signal generators, valve voltmeters, frequency monitors, counters, waveguide test equipment and other products of the Hewlett-Packard Company of California, U.S.A. Technical service in this country provided by Livingston Laboratories. A short-form illustrated catalogue from the British representatives, Lithgow Electronics, 1, Grange Court, Sudbury Hill, Harrow, Middlesex.

Wide-Band Television Aerials

Review of the More Interesting Types Current in North America

By M. G. O'LEARY

THE recent inauguration of television in Band III has introduced at least one problem at the receiving site—that of an aerial system which will efficiently receive signals in both bands. Preferably this system should bring in both bands on a single transmission line. British manufacturers have introduced numerous designs to do just this. In the light of these events, a review of practice in North America, where television on both bands is several years old, should be of interest to students of aerial design.

There are several differences between British and North American television and since the reader is presumably familiar with British standards, only the American ones need be reviewed. Channels are 6 Mc/s wide. Channel 2, the lowest, is 54 to 60 Mc/s, Channel 3 is 60 to 66 Mc/s, and so on to Channel 6 at 82 to 88 Mc/s, with a 4-Mc/s gap between Channels 4 and 5. Channels 7 to 13 on the high band cover 174 to 216 Mc/s. Other differences include transmission of 30 pictures per second with 525 lines. Nevertheless, these differences do not alter the basic aerial problem for Band I/Band III reception. However, in North America television signals are transmitted with horizontal polarization and this difference presents an additional consideration to the aerial designer, but only with respect to structural and mechanical factors. A good horizontally polarized aerial is generally a good vertically polarized one if it is rotated 90 degrees about its axis. Therefore it follows that a successful American aerial design for dual-band reception will also be successful for this in Britain, if mechanically and structurally it is feasible to rotate it through 90 degrees. Nevertheless, American designs which do not meet this requirement will be reviewed here also if their design and theory are unusual. Ingenuity will always suggest adaptation in one form or another.

The basis of the problem is the three-to-one (approximately) frequency ratio between Band-I and Band-III channels. A half-wavelength dipole cut for a Band-I channel will be one-and-a-half wavelengths for Band-III channels. The polar response and current distribution of this dipole for the Band-I channel are as in Fig. 1, whereas these characteristics for the same dipole on Band III are shown in Fig. 2. The split lobes in Fig. 2 could be used for reception of horizontally polarized signals, at the risk of ghosts, but such an aerial would give extremely poor results on vertically polarized Band-III signals, even when rotated through 90 degrees. Nevertheless, no American designs known attempt to receive Band-III frequencies in this way. Instead, an attempt is made to alter the Band-III polar pattern to obtain a single forward lobe in line with the lobe for Band I.

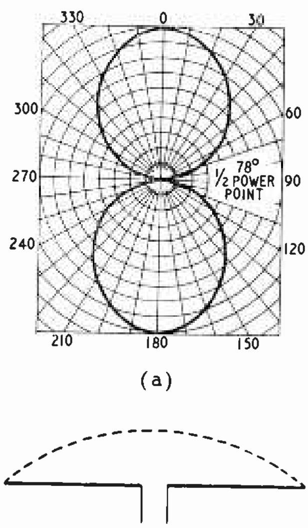
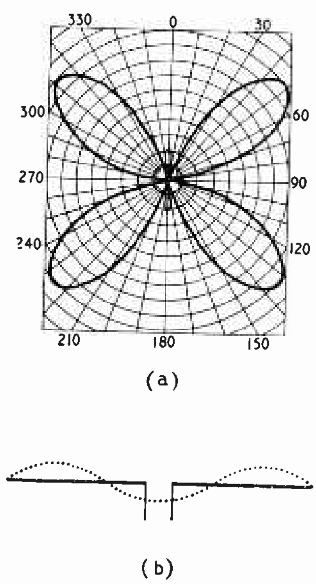


Fig. 1. Polar diagram (a) and current distribution (b) of a half-wave dipole cut for Band I.

Fig. 2. Polar diagram (a) and current distribution (b) of a Band-I dipole used as a 3/2 wavelength aerial on Band III.



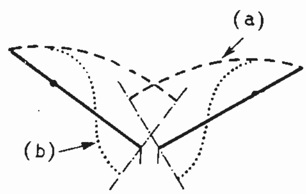


Fig. 3. Current distribution on a tilted-element aerial; (a) Band I (b) Band III.

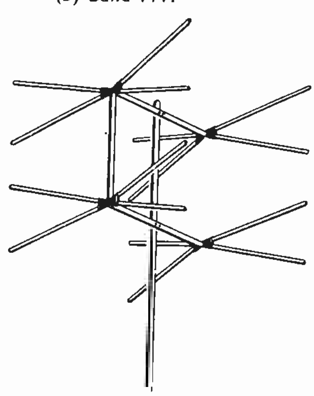


Fig. 4. Two forward tilted conical dipole aerials (with reflectors) stacked vertically for fringe areas.

A widely used means of accomplishing this is to cut a Band-I dipole and tilt its elements forward as in Fig. 3. Doing this does not adversely affect the Band-I polar pattern, but it combines the split lobes of the Band-III pattern into a single lobe. In doing so, small split lobes remain in the opposite direction, which could cause ghosting troubles in difficult localities. Usually these tilted aerials have broad-band, conical-type dipoles with a reflector favouring the Band-I frequencies. In fringe areas two are stacked vertically with a spacing favouring Band III, as in Fig. 4. Such stacked arrays give excellent reception at distances of 75 miles from transmitters on both bands (100 kW e.r.p. Band I; 200 kW e.r.p. on Band III).

Another design popular some years ago functions on a different principle (see Fig. 5); A is a half-wavelength Band-I dipole, B is a full-wavelength Band-III dipole, "T"-matched to a balanced transmission line through C. On Band I, A is doubly "T"-matched to the transmission line through B and C, but the connections between A and B are low-pass filters acting as a short circuit on Band I, and as an open circuit on Band III. Thus A is not directly effective on Band III other than to act as a band broadening device through mutual coupling to B; B in turn acts as a broadening device on A for Band-I frequencies. These aerials were supplied with a reflector effective on Band I and a director favouring Band III, and were similar to the tilted conical dipole and reflector in performance.

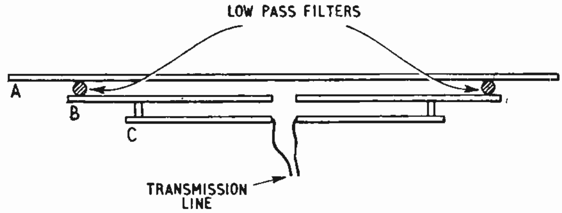


Fig. 5. Band I/Band III combination aerial embodying low-pass filters for coupling elements. (Lapointe Electronics Co.)

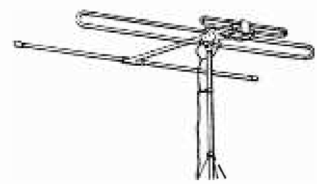


Fig. 6. In-line combined aerial with two folded dipoles.

Collinear Designs

The "inline" aerial, Fig. 6, consists of two folded dipoles, one for Band I behind the other for Band III. The larger acts as a reflector for the Band-III frequencies and, in turn, is itself aided by a reflector cut for Band I. The two dipoles are served by a single transmission line, which, along with the dipole spacing, is so arranged that there is a minimum of interaction between the dipoles when acting on their respective frequencies.

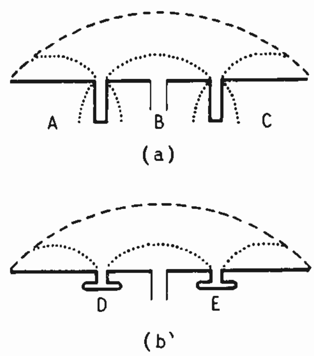


Fig. 7. Collinear arrangement of dipoles with (a) closed-end connecting stubs and (b) folded stubs. Current distribution shown also. Dashed-line Band I, dotted-line Band III.

Collinear designs as in Fig. 7(a) and (b) have also been popular. In Fig. 7(a), A, B, and C are each a half-wavelength for Band III, separated by closed-end quarter-wavelength stubs (Band III). The current distributions for each band are indicated (compare with Fig. 2) and the result is a single-lobe polar response on both bands. A popular configuration takes the form of four such aerials stacked vertically and backed by a screen or reflector spaced to give good reflector action on both bands. Referring again to Fig. 7(a), the total length A+B+C, and the separating stubs, is effective for Band-I response. Stacking spacing of these multiple aerial systems is a half-wavelength on Band III, and thus favours this band.

Figure 7(b) illustrates a different way of accomplishing this collinear effect. Rather than separate the three collinear elements with quarter-wavelength stubs, small quarter-wavelength (Band III) folded dipoles, D and E, are used. These reverse the phase of the Band-III currents by 180 degrees and thus all three collinear elements act in phase to give a single-lobe polar response. On Band I the combination acts as a single dipole. Fig. 8 shows how this design can be used for a two-stack yagi system.

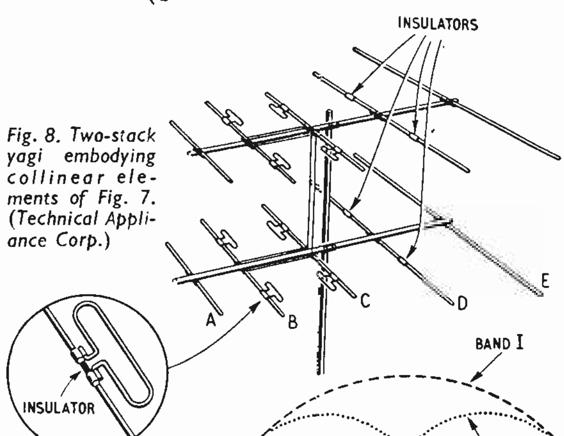
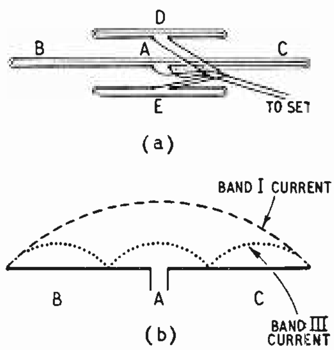


Fig. 8. Two-stack yagi embodying collinear elements of Fig. 7. (Technical Appliance Corp.)

Fig. 9. Current distribution on director A of Fig. 8.



Left: Fig. 10. Triple folded dipole aerial (a) with current distribution (b).

Figure 8 calls for some additional explanation—

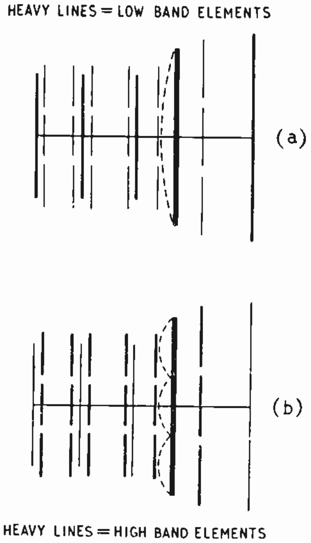


Fig. 11. Current distribution on Bands I and III of triple folded dipole aerial of Fig. 10 with added parasitic elements.

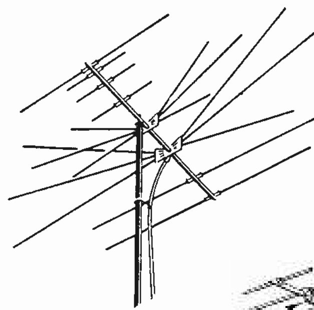
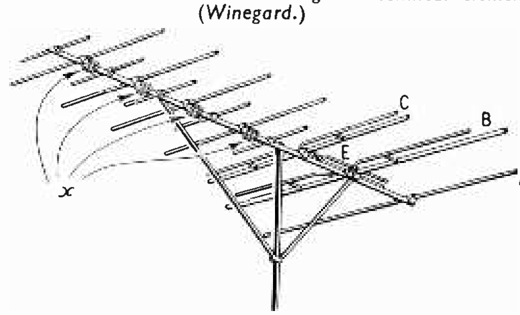


Fig. 12. In-line design using conical dipoles on both bands and several parasitic elements. (Kay-Townes Antenna Co.)

Below : Fig. 13. Dual-band yagi using "T"-matching and collinear elements. (Winegard.)



the "fed" elements (those to which the feeder is joined), are spaced a quarter-wavelength for Band I and three-quarter-wavelength for Band III. Twin-fed dipoles spaced a quarter-wavelength and fed at the rear are well known for having a good front-to-back ratio (not so true when vertically polarized), broad bandwidth, and fairly good sensitivity.[1] Reasonably similar results could be expected on Band III with three-quarter-wavelength spacing. Director A is actually two collinear Band-III directors separated by a quarter-wavelength transmission stub (Band III), the whole acting as a Band-I director as well. Reflector E is cut for Band I. Reflector D is three collinear Band-III reflectors insulated from each other by fibreglass rods. D has no significant effect at Band-I frequencies; E is spaced too far from C to be effective at Band-III frequencies.

Figure 9 illustrates the action of the director A in Fig. 8, and shows the current distribution for the two bands. This idea has been used also for stacked dipoles where both elements are fed and backed by a screen-type reflector. The dipoles are made of angle section for structural as well as wide band considerations and slotted to reduce wind resistance. The stacking distance favours the Band-III channels, while the reflector spacing favours the Band-I frequencies.

Figure 10 (a) and (b) illustrate the development of the so-called "tripole" type of aerial. Two Band-III folded dipoles are coupled with a single Band-I folded dipole and connected together so that dipoles D and E not only neutralize the out-of-phase Band-III current in the A section, but provide a resultant current from the combination D, A, and E which is in phase with those at B and C. D and E have negligible effect at Band-I frequencies. This "tripole" has been used both with screen-type reflectors, and in yagi configurations. In the latter form directors and reflectors for both bands are interspaced with the other elements. Those for Band III are three collinear elements isolated from each other with fibreglass couplers. Fig. 11 (a) and (b) illustrate the action of the various elements on the respective bands.

A number of more complicated "inline" designs have been introduced recently which are alleged to be much more sensitive than the simpler designs, such as the one illustrated in Fig. 6. Fig. 12 shows a typical example. The larger conical dipole receives Band-I signals while the smaller catches the Band-III signals. The inter-connections between the two, in combination with spacings and associated parasitic elements, neutralize the normal split lobes of the conical dipole on Band III and shape a single forward lobe. This "inline" design has a half-wavelength conical dipole for Band I behind a full-wavelength one for Band III.

Complex Systems

An extremely interesting type of yagi has recently been introduced based on a somewhat different principle to any previous designs. Fig. 13 illustrates the use of this principle in a dual-band yagi design. Since the theory of this yagi is a combination of this and several other principles, some elaboration seems indicated. On Band I the active portion, A, B and C, is a twin-fed yagi, stagger-tuned to give better cover of the entire band. The forward dipole, C, is connected to the transmission line, and the inter-connection, E, to the rear dipole is transposed. The spacing between the two dipoles is one-eighth wavelength for the median frequency of Band I. This configuration gives a good front-to-back ratio, broadband characteristics, and, for normal dipoles, very low aerial feed-point resistance.[2] This latter property, normally not desirable, is countered with a specially designed "T" match (or dipoles B and C) to raise the resistance. Incorporated into the "T" match are two collinear half-wavelength dipole elements for Band III reception. The three-eighths-wavelength spacing which results for Band-III frequencies is not quite as effective as the one-eighth-wavelength spacing of Band I, but the large number of parasitic elements on Band III more than make up for this. Close observers will discern that the long directors are, in principle, the same design as those in Figs. 7 and 8, but here the quarter-wavelength shorted transmission stubs, x (for Band III), are folded for compactness.

The designs discussed so far have all attempted

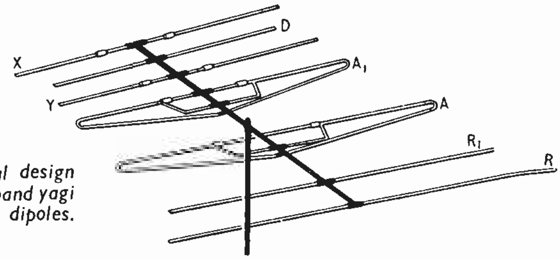


Fig. 14. Unusual design of in-line, dual-band yagi using "wing" dipoles. (Trio Mfg. Co.)

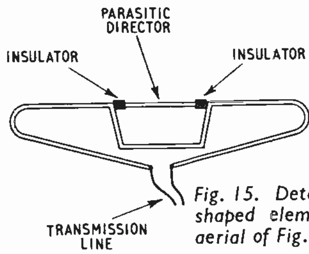


Fig. 15. Details of wing-shaped elements in the aerial of Fig. 14.

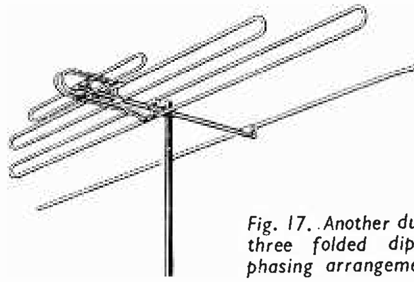


Fig. 17. Another dual-band system with three folded dipoles, and unusual phasing arrangement (Sabre).

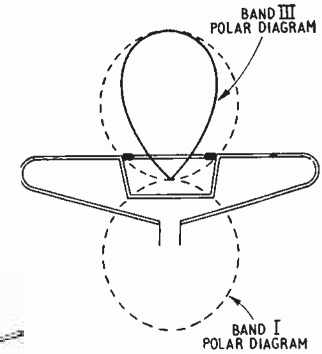


Fig. 16. Polar diagrams of "wing" dipole shown in Fig. 15 on Bands I and III respectively.

to obtain single forward response lobes on both bands exclusive of parasitic action; that is, in the primary (fed) elements only. Lately several designs have been introduced in which parasitic action has been coupled with primary element characteristics to accomplish the same result. One such is shown in Fig. 14. Here Band-I action is that of a stagger-tuned, twin folded-dipcle (A, A_1) yagi with one director (D) and two reflectors (R, R_1). Fig. 15 illustrates the configuration of the wing-shaped folded dipoles. Two of the three directors (X, Y) are actually triple collinear Band-III directors as discussed for previous designs. Fig. 16 shows how the dipole configuration, plus the built-in parasitic director, act together to produce a single forward lobe for Band-III frequencies.[3] There are other such designs which allegedly correct Band-III directivity parasitically.

All the designs discussed are claimed to be broad-band types having good aerial characteristics on all twelve channels. The writer has witnessed that this is so for a good many of them in an area where the following channels could be received: 2, 150 miles; 3, 75 miles; 4, 150 miles; 5, 80 miles; 7, 40 miles; 8, 75 miles; 10, 80 miles; 11, 16 miles; 12, 160 miles. Extending the responses of an aerial to cover the high-band channels is not excessively difficult, but the low-band channels are not quite so easily covered as the frequency ratio is wider, the band being 54 to 88 Mc/s. In some of the designs discussed the broadening devices are quite apparent and in general these take the form of conical-type dipoles, stagger-tuned dipoles, director lengths favouring the high end of the band, reflector lengths favouring the low end of the band, parasitic spacing favouring one or another portion of the band, twin yagis, large diameter dipoles, folded dipoles, etc. In other types the broad-band devices are not so obvious; apparently element interaction and mutual coupling having been used in these cases to give the desired results. Also all these designs are claimed to be a good match on all 12 channels to the 300-ohm transmission line which seems to be the most commonly used type in North America.

No attempt has been made in this discussion to include all of the multiplicity of dual-band aerial designs available in North America. Emphasis has been placed on principles of operation rather than on the many variants of a given principle that are available. But, before closing, the writer cannot resist the temptation to include Fig. 17, a dual-band design with claimed "Miracle Phase." Not having seen the patent, or other explanation of its theory, the writer hesitates to put forward his analysis of its action. It is presented mainly as a bit of mental exercise for students of aerial design.

REFERENCES

[1] *T.V. and Other Receiving Antennæ*, A. B. Baily (Rider).
[2] *The A.R.R.L. Antenna Book.*
[3] *Radio and Television News*, October, 1955, p. 91

"Analysis-Synthesis" Telephony

ECONOMIES in bandwidth of the order of 100:1 are envisaged in a system of speech transmission, under development by the Post Office Research Station, which was demonstrated at the Royal Society Conversazione in May.

Speech sounds can be synthesized by applying pulses of different amplitude and repetition rate (larynx excitation) to resonant tuned circuits (cavity formants of the mouth, etc.) and adding bursts of white noise (hissing consonants). When circuit elements of this kind are connected to a loudspeaker and energized in the proper sequence by signals originating from an equivalent analysis of the speech at the sending end, intelligible and often realistic speech is heard.

In the Post Office analyser three formant frequencies were selected by tuned circuits, the larynx tone by isolating the peaks, and the hissing sounds (fricatives) by their high-frequency content. Some ambiguities are inevitable in a simple analyser of this kind, but the results so far achieved are undoubtedly promising.

For convenience a six-way cable was used in the demonstration to connect the transmitter and receiver, but there is no reason why the information should not be encoded for transmission on a single channel.

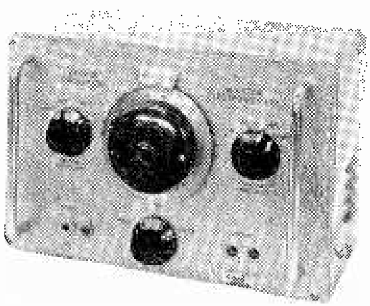
"Magnetic Tape Amplifier." In Fig. A, p. 123 of the March issue a 0.1μF blocking capacitor should be included in the lead from the junction of R_{10a} and R_{10b} to the twin-T network.

Manufacturers' Products

NEW EQUIPMENT AND ACCESSORIES FOR RADIO AND ELECTRONICS

Decade Oscillator

BOTH sine and square waves over a frequency range of 10 c/s to 100 kc/s are provided by the decade oscillator made by Winston Electronics, Ltd., Govett Avenue, Shepperton, Middlesex. It is of the Wien bridge type with thermistor amplitude stabilization



Winston Electronics decade oscillator, 10 c/s-100 kc/s, sine or square waves

within 1%. The output attenuator is calibrated 0-10V with switched multiplier of ×0.1, ×0.01 and ×0.001.

Harmonic distortion on sine waves is said to be <1% and on square waves the rise time is about 0.3 μsec at 100 kc/s. At 10 c/s the maximum drop in the horizontal part of the wave is 2%. Frequency stability is <1% for ±10% change of mains voltage and about 0.02% for ambient temperature changes.

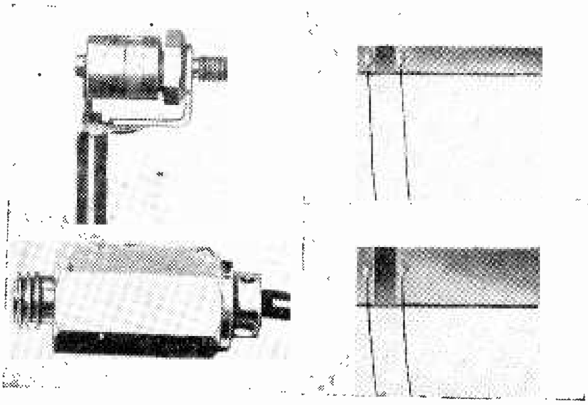
Valve replacement is simplified by the use of 12AT7s throughout.

The price of the decade oscillator is £57.

Barium Titanate Transducers

AN accelerometer and strain gauges for vibration testing are now being produced by the General Electric Co., Ltd., Kingsway, London, W.C.2, in which piezo-electric barium titanate elements perform the conversion from mechanical strain to electrical output.

The accelerometer (Type E) makes use of a disc element in contact with a 10-gram mass. An alternating acceleration of "*g*" (equivalent to the acceleration



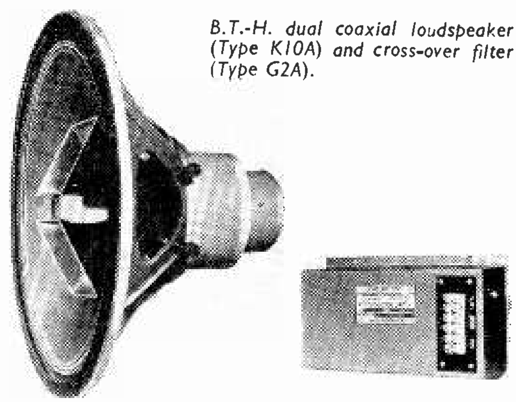
(Left) G.E.C. barium titanate accelerometer in unscreened and screened versions, and (right) vibration strain gauges

due to gravity) gives an output of about 20mV and this is maintained within ±10% over a range of 40 c/s to 10 kc/s. The transfer characteristic is stated to be linear up to 1000*g*. To minimize spurious readings the transverse sensitivity has been limited to 5% of the axial sensitivity.

Like the accelerometer, the strain gauges are intended for alternating displacements (originally, vibration in turbine blades) and it is claimed that their sensitivity is over 2,000 times higher than conventional resistance elements. They are 0.035-in in thickness, ¾-in long and either ¼-in or ⅛-in wide; the frequency range quoted is 20 c/s to 50 kc/s. An important advantage is that they can be used as driving elements to excite as well as detect resonances. Temperature limits are −50°C to +100°C.

Dual Moving-Coil Loudspeaker

THE B.T.-H. Type K10A moving-coil unit used in cinema sound installations can now be bought for use with high-quality domestic equipment. Frequencies up to 1700 c/s are radiated from an 18-in cone and above that frequency from a coaxial horn-loaded pressure unit.



B.T.-H. dual coaxial loudspeaker (Type K10A) and cross-over filter (Type G2A).

After emerging from the centre pole of the large magnet the horn divides into twin flares, arranged to maintain wide distribution in the horizontal plane at high frequencies.

The l.f. drive unit has a magnet with a flux density of 14,300 gauss and total flux of 285,000 maxwells; the figures for the h.f. unit are 12,700 gauss and 48,000 maxwells. The power handling capacity of the unit as a whole is rated at 20 watts.

Particular attention has been paid by the makers to the delayed resonance response, and this is claimed to be free from anomalous effects.

A two-section cross-over filter (Type G2A) with an attenuation of 12dB/octave at cross-over is included in the price of £45. The makers are the British Thomson-Houston Co., Ltd., Rugby.

We regret publication of this issue of *Wireless World* has been delayed. The next issue (July) should be dispatched on June 29th, but with the August number we hope to resume normal publication (on the fourth Tuesday of the preceding month).

Aerial Cross-Over Network

Design and Construction of a Unit for Combining Band I and Band III Aerials

By L. S. KING, B.Sc.(Eng.), A.M.I.E.E.

THE cross-over network is an electrical filter unit which enables Band I and Band III aerials to be coupled together to a common downlead to the television set. It consists of precise values of inductances and capacitances to meet certain known frequency and impedance data.

The advantage of such an arrangement lies in the single downlead, and the ability to select either Band I (B.B.C.) or Band III (I.T.A.) by simply turning the knob of the tuner on the set. Of course, where a combined Band I/Band III aerial is used, the cross-over network is not required and the said advantage does not arise—or for that matter perhaps, where the set (or set and convertor) has two inlet sockets.

A disadvantage in the use of the cross-over network is not operational, but electrical. The introduction of any network must involve electrical losses because the network components are not pure reactances (which alone would be non-dissipative) and the position that arises is whether any losses at all in signal strength can be afforded. On Band I, in the service areas, some loss can usually be accepted, but this state of affairs is often not the case on Band III with its considerably higher radio frequency and greater attenuation in signal strength.

The two aerial systems to be coupled may individually be producing pictures on I.T.A. as brilliant and as clearly defined as on B.B.C. and with a background *almost* as steady, but the condition inside the receiver circuitry may not be similar and may be very different in the two cases. The receiver is provided with automatic gain control (a.g.c.), and this will attempt always to match up the signal to the required level by a varying amount of amplification, but to this, of course, there is a practical limit. If the I.T.A. signal seems poor and on looking carefully at the picture there is a faint stirring of the background as compared with a steady background on the B.B.C. channel, then it might be wise to leave well alone and not introduce any further losses at present on the I.T.A. channel. However, there is no reason why such a filter should not be tried as it is relatively easy and cheap to construct; also, it may be possible to improve the signal strength by means of a larger and higher aerial.

The network is an electrical filter, although hardly a filter as would be recognized as such in the telecommunication field, where steep-sided filters are enforced by the closeness of adjacent communication channels and the necessity of non-interference between these close-spaced channels. In the case of Band I/Band III, we have something like 140 Mc/s separation between the relatively low frequencies of Band I and the higher frequencies of Band III. Consequently only a few inductors and capacitors are required in the relatively simple filter which can be used, since steepness of cut-off is not a necessity, and a much more gentle slope up to cut-off will suffice. The filter will then have two distinct portions, one called the low-pass (LP) section which will allow the passage of Band I frequencies with low loss, but which will offer a high impedance to Band III frequencies so as effectively to cut off the latter. The other portion, called the high-pass (HP) section, will allow the passage of Band III frequencies with low loss, but will effectively cut off Band I frequencies.

With these points established, there is no reason why the relatively slow rising characteristic curve of each portion should not overlap as shown in Fig. 1, the only necessary condition being that each frequency in turn is passed by its own filter section and not by the other.

This allows the use of the simplest form of filter and, in general, of one that would be useless in the telecommunication art; that of the half section as shown in Fig. 2.

It will now be seen from the general shape of the characteristic curves of this type of simple filter

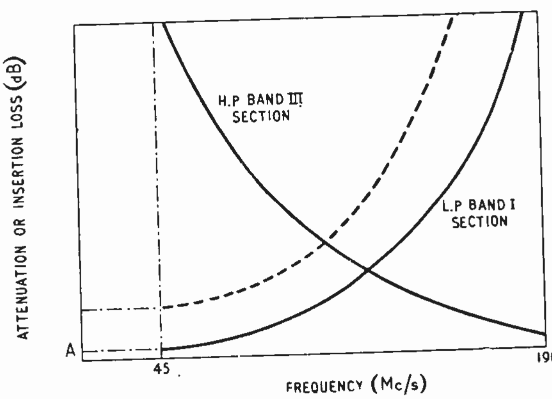


Fig. 1. Attenuation, or insertion loss, of the high-and-low-pass filter sections comprising the aerial cross-over network described in the text.

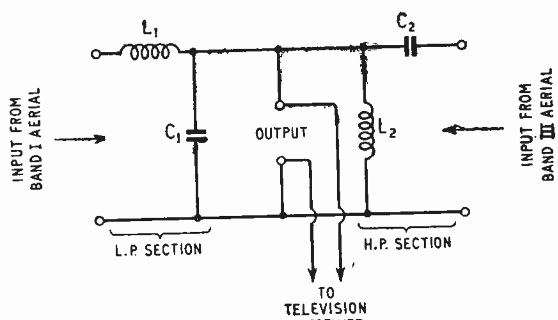


Fig. 2. Theoretical circuit diagram of the aerial cross-over network.

section, Fig. 1, that in the case of the LP section (Band I), say, the cut-off of that section must be well up to the Band III operational frequencies for the rather gentle slope of the curve to produce a sufficiently low attenuation, A, at its own operational frequencies. Imagine a curve of the same family drawn more to the left and shown dotted in Fig. 1. It would still be effective to Band III frequencies as it would cut off at an even lower frequency, but the attenuation, or insertion loss, at Band I frequencies has now gone up from A to A′. This means that the insertion loss at its admittance frequency has probably increased to a prohibitive value and so the filter would be unsuitable for our purpose. Similarly, the HP section of the filter should be designed for a cut-off frequency near to Band I frequencies, and here, low insertion loss to Band III frequencies is even more imperative.

Having determined on this, as the basis of design, we will set the cut-off of the LP filter section at 190 Mc/s and that of the HP filter section at 50 Mc/s, both to match into 70 ohms impedances.

Formulæ for the calculation of filter components are given in the Appendix, and using these, we get the values for capacitance C shown in the following table:—

CAPACITOR TABLE

Position	Calculated Value (pF)	Use as below (pF)
C_1	24	25
C_2	23	50 + 40 in series = 22.2

Now it is not usually possible to get capacitors with capacitances as calculated and some compromise has to be made. Also, some series or parallel arrangement may have to be built up; an expedient that is adopted here.

For example, 23 pF is made by putting 50 pF in series with 40 pF to give 22.2 pF as the nearest to 23 pF. Remember the rule for adding capacitors in series is similar to resistors in parallel, viz.;

$$\frac{\text{Product}}{\text{Sum}}, \text{ or } \frac{50 \times 40}{50 + 40} = \frac{2,000}{90} = 22.2$$

Low loss silvered-mica capacitors should be used. Making the inductors is liable to give trouble owing to the fractional number of turns as calculated and the small diameter of the former that is required. The writer found that while 4 turns on L_1, produced a fair picture, 3.8 turns gave a better one, so the inductance apportioning is fairly critical. Formulæ for simple-layer inductors are well known and using No. 26 s.w.g. insulated wire on $\frac{1}{4}$in. dia. polystyrene formers the number of turns required are given in the inductor table. The wire should be secured to the formers with a small blob of "Styrene" cement (polystyrene dissolved in a little

INDUCTOR TABLE

Position	Inductance (μH)	Turns on $\frac{1}{4}$in dia former
L_1	0.12	3.85
L_2	0.112	3.75

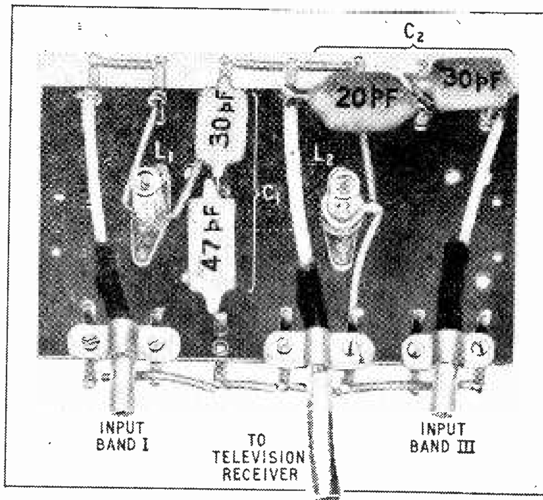


Fig. 3. Practical version of the aerial cross-over network. The thin laminated base measures $2\frac{1}{2} \times 4\frac{1}{2}$in and should be mounted in a screening box about 2in deep

benzine). The stock type formers are internally threaded for dust cores but no cores are used with these coils.

Obviously, it will not be possible to achieve these fractional portions of a turn but they are given here so that the constructor knows what should be aimed at. It is not possible to achieve 3.85 turns but 3.8 turns can be obtained and so a compromise is made.

Final Design

Distributed capacitance in the coils will be very small but it is still appreciable when dealing with extremely small values. This capacitance would have the effect of lowering the impedance to something lower than the expected impedance as calculated. It is extremely difficult to get an idea of the capacitance due to the coils, so it would be well to design for an impedance somewhat higher than 70 ohms, say of the order of 85 to 90 ohms. For a nominal 90 ohms impedance, the component

L AND C VALUES FOR 90-OHM IMPEDANCE FILTERS

Section	Component	Value	How Derived
LP	C_1	18.6 pF	30 pF and 47 pF capacitors in series = 18.3 pF nearest value.
	L_1	0.15 μH	$4\frac{1}{2}$ turns No. 26 s.w.g. double silk covered wire close wound on $\frac{1}{4}$in dia polystyrene former
HP	C_2	12.6 pF	20 pF and 30 pF capacitors in series = 12 pF nearest value
	L_2	0.102 μH	4 turns No. 22 s.w.g. double cotton covered wire close wound on $\frac{1}{4}$in dia polystyrene former

Capacitors to be $\pm$ 5% tolerance or better.

values are as follows, but here the wire gauge for L_2 has been changed to No. 22 to obtain an even number of turns for the inductance required.

The writer has made up networks to both 70 and 90 ohms but the nominal 90-ohm network was preferable. This network is illustrated in Fig. 3 and is shown assembled on a thin laminated base with tag spacings of ⅝in. and 2in. centre-to-centre across. Judging from the B.B.C. picture there was no evidence from the brightness or definition that the filter had been inserted, but on Band III where, in the writer's case, little signal strength can be sacrificed, the insertion of the filter did produce some reduction in brightness, but generally to the extent that it was noticeable only when compared with the brightness with the filter removed.

APPENDIX

The filter types used are constant-*k*, half-section networks, and the formulae used for these half-sections are as follows:—

Low Pass Section $L_1 = \dfrac{R_o}{\pi f_c}$

$$C_1 = \frac{1}{\pi f_c R_o}$$

High Pass Section $L_2 = \dfrac{R_o}{4\pi f_c}$

$$C_2 = \frac{1}{4\pi f_c R_o}$$

where L is in henrys
C is in farads
R_o is the equivalent terminating resistance in ohms
f_c is the relevant cut-off frequency in c/s.

Abnormal V.H.F. Propagation

Determination of Radio Refractive Index Structures from Weather Data

By A. H. HOOPER

IT is now well known that there is a considerable degree of association between v.h.f. propagation and weather conditions. On the seasonal scale, for example, the general level of signal strength is weaker in winter than in summer, while over periods of several days duration marked departures from seasonal averages are found to develop. On the latter occasions it is frequently found that a spell of fine settled weather is being experienced. On a still smaller time scale, signal strengths are found to increase and decrease from normal for a matter of hours

In time these effects come to be regarded as associated with the weather conditions observed. Detailed examination has revealed, however, that such associations exist in only a proportion of cases; there are, for example, many spells of fine settled weather with nothing unusual in the way of propagational effects occurring. In consequence assessments of propagational conditions cannot usefully be made from studies of local weather conditions or of weather charts alone.

It has been found that fluctuations of signal strengths at metre wavelengths can, to a very large extent, be explained by variations in the amount of downward bending of the radio waves in passing through the lower levels of the atmosphere. The significant quantity in such circumstances is the vertical structure of radio refractive index. The only data from which this can be determined on a day-to-day basis are the results of meteorological soundings of the atmosphere given in the form of a series of values of air pressure, dew point and temperature.

On a day of vigorous atmospheric motion it is often found that the radio refractive index decreases with height at a fairly steady rate and that signal strengths are sub-normal. On occasions of stagnant atmospheric conditions, however, large departures from a steady rate, and with them enhanced signal strengths, may occur. While the former condition can, with practice, be ascertained by inspection of the meteorological sounding data it has been found necessary both for the latter case and for the more usual case of intermediate conditions to determine approximately the vertical structure of radio refractive index. Precise evaluation of radio refractive index values for quantitative work is necessary only on the proportion of occasions when significant variations in structure occur.

The method to be described enables a graph of radio refractive index with height to be prepared very rapidly from meteorological data. The state of the lowest layers of the atmosphere is clearly displayed, and on those occasions when numerical values of radio refractive index and height are required, use of a measuring scale enables them to be read with as much accuracy as the basic observations justify. As the numerical computation of a series of values of radio refractive index from the

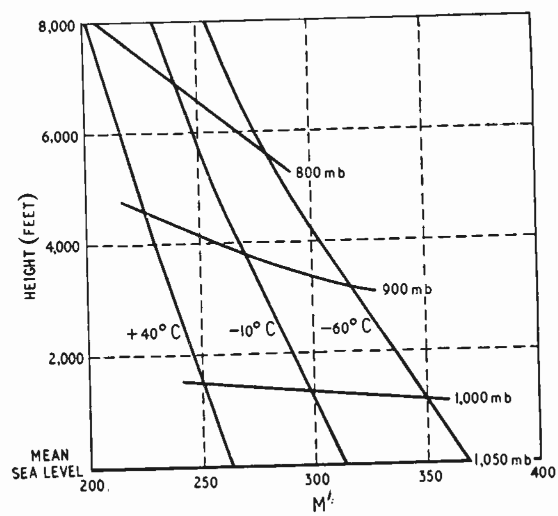


Fig. 1. Elementary form of graph giving a solution for the first term of the basic equation.

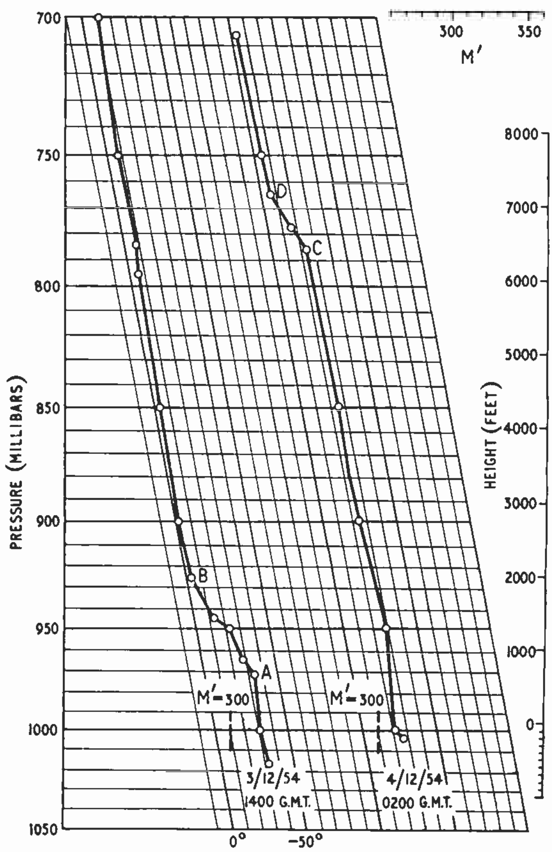


Fig. 2. Plotting chart in terms of pressure and the "refraction temperature" derived from Fig. 3.

reported data is a laborious and time-consuming task it is evident that the proposed method offers considerable advantage.

Derivation of Graphs.—The radio refractive index μ is often expressed in "M-units", given by

$$M' = (\mu - 1).10^6$$

This yields values ranging in the lower troposphere from 360 to 280 units.

In terms of meteorological parameters, M′ is approximately given by

$$M' = \frac{79.P}{T} + \frac{379200.e}{T^2}$$

where P=total atmospheric pressure (millibars), e=(partial) water-vapour pressure (millibars) and T=temperature (degrees absolute). The error in M′ is less than 1% at below 336°A (=63°C).

The graph adopted for displaying radio refractive index structure has height as ordinate and M′ as abscissa, both being linear and increasing conventionally.

A graph of this type is shown in Fig. 1, together with superimposed curves of P and T derived from the expression $\frac{79.P}{T}$. A plot of observed values of P and T using the superimposed grid gives, upon reference to the underlying grid of M′ and height, a direct solution for the first term of the given equation. For the rapid evaluation of the complete equation, however, there are advantages in replacing these superimposed lines by a sufficiently accurate approximation in the form of Fig. 2, from which the basic grid of M′ and height has been omitted for clarity. This alternative grid comprises a set of horizontal pressure lines and straight, parallel, temperature lines. The magnitude of the approximation is considered later.

(Continued on page 297)

TABLE I

Distance, in inches, of pressure lines in Fig. 2 above datum (1050 mb)

Pressure Millibars	0	10	20	30	40	50	60	70	80	90
700	11.02	10.63	10.25	9.88	9.51	9.14	8.78	8.43	8.08	7.73
800	7.39	7.05	6.72	6.39	6.06	5.74	5.42	5.11	4.80	4.49
900	4.19	3.89	3.59	3.30	3.01	2.72	2.43	2.15	1.87	1.60
1000	1.33	1.06	0.79	0.52	0.26	0				

TABLE II

Values of Tr

	Dew Point (°C)										
Air Temperature (°C)	Absolute zero	−20	−10	0	+5	+10	+15	+20	+25	+30	+35
+40	+40	+35	+29	+16	+6	−7	−26	−50	−83	−124	−178
+30	+30	+25	+18	+5	−6	−21	−40	−67	−101	−145	
+20	+20	+14	+7	−7	−19	−34	−55	−83			
+10	+10	+4	−4	−19	−31	−48					
0	0	−6	−15	−31							
−10	−10	−17	−26								

The second term of the equation can be obtained by a graph relating e and T and the position in Fig. 2 denoting total M′ value then found by moving, from the position given by the first term, through the corresponding number of units to the right. Since the expression for the second term, $\frac{379200.e}{T^2}$, does not contain P this horizontal distance is constant for all pressure levels. It can, therefore, be expressed in units of the parallel temperature lines. Hence the final position denoting M′ can be found by the intersection of the appropriate pressure line, and a "temperature" line obtained by adding to the observed temperature a certain number of degrees determined separately by a graph of T and e. This leads to the concept of an "effective refraction temperature" T_r which gives directly the final sloping temperature line required.

The graph of Fig. 3 combines the above steps to give directly the refraction temperature corresponding to each combination of air temperature and dew point likely to be experienced in normal work. The process of deriving the position of radio refractive index value is reduced, therefore, to ascertaining T_r from Fig. 3 and plotting a point on Fig. 2 at the intersection of the corresponding sloping T_r line with the appropriate pressure line.

Numerical Values.—In most cases the structure given by a series of points so determined will be all that is required. When, however, the approximate height of a significant point is required it can be directly measured by placing a straight edge, graduated to the basic height scale adopted, between the pressure levels of the surface and of the point concerned. Values of radio refractive index for a given occasion are conveniently obtained by drawing a vertical line representing the value $M'=300$ through the intersection of the $-10°C$ temperature line and the 1000 mb pressure line and then measuring with an appropriate scale the horizontal distances from this line to the series of plotted points. The scale is graduated in M′ values and when placed in registration with the line gives a direct reading of M′.

Specifications.—The results of soundings of the atmosphere made twice daily at stations of the Meteorological Office are published in the *Daily Aerological Record**. The information is in the form of temperature and dew point at pressure levels selected individually for each sounding so as to delineate the observed structure. On 1st January, 1956, the scale of temperature was changed from Fahrenheit to Centigrade, and the chart and graph are designed for use with the current temperature scale.

For the plotting chart (Fig. 2) it is recommended that one inch represent 1,000 feet and 50 M-units. It is then possible to use a rule graduated in inches to read both M′ and height. Table I gives the distance from the lower edge at which the pressure lines are drawn.

The T_r lines are drawn, five to the inch, with a slope of -1.9 inches in ten inches of height. The given separation arises from the convenient fact that, between $-10°C$ and $+30°C$, one degree Centigrade is closely equivalent to one M-unit.

* Obtainable on application to the Director, The Meteorological Office, Dunstable, Beds. Single copies 4d, 1 month 9s, 3 months 24s, 1 year 95s.

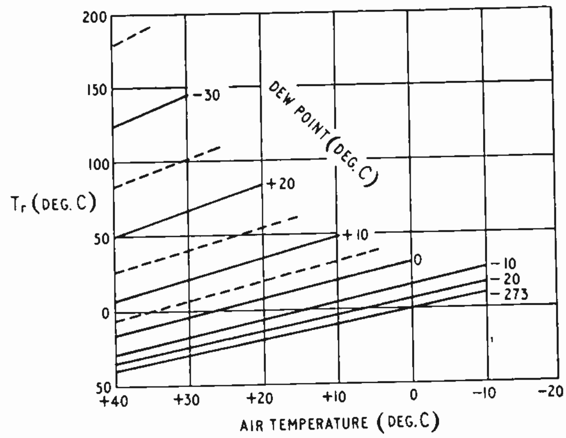


Fig. 3. Graph from which "refraction temperature" can be derived from observed values of dew point and temperature at different pressure levels.

TABLE III
Chart error (M′ units)

Pressure (mb)	Temperature (°C)				
	+30	+20	+10	0	−10
700... ...	−12	−9	−6	−3	0
800... ...	−5	−3	−1	+1	+3
900... ...	−2	−1	0	+1	+2
1000... ...	0	0	0	0	0

When much work is contemplated the chart is best prepared on tracing linen and then duplicated as required.

Table II gives data for reproducing the T_r graph of Fig. 3.

Accuracy.—Considering Fig. 1 it will be seen that replacement of the converging curved temperature lines with a set of parallel straight temperature lines introduces an error in M′ which varies with position in the diagram. The slope selected for the straight lines has been chosen to minimize this error in the normal area of use. The magnitude of the error is given in Table III. It will be found that errors in excess of three units are extremely rare.

There is, of course, additional uncertainty arising from errors in reading from the graph of Fig. 3 and from plotting.

The height at which a point is plotted will be in error by a factor related to the departure of the mean temperature of the air column (between surface and the level) from the mean temperature (283°A) assumed in Fig. 2. Correction is at the rate of 7% for every ten degrees of departure. On nearly all occasions in the vicinity of the British Isles the correcting factor is less than 3%.

In making use of the radio refractive index structure obtained from radiosonde data it is assumed that the given values are representative of an area sufficiently large for the purpose in hand. It is necessary, therefore, to pay attention to the effect both of instrumental uncertainty and of atmospheric inhomogeneity. Consideration of these effects leads to the conclusion that the standard deviation of a spot value of radio refractive index from the British radiosonde is about 10 M-units, while the uncertainty of the height of a given point is represented by a standard deviation of 217 feet. It is concluded, therefore, that on most occasions the approximations of Fig. 2 are acceptable and that the method is sufficiently accurate to depict meteorological sounding data in this form.

Example.—The radio refractive index structures derived from the results of two successive soundings over Sussex in December, 1954, are shown in Fig. 2. A marked zone of discontinuity can be seen, between A and B on the earlier and between C and D on the later sounding. From these, and the results from adjacent areas, it is apparent that it is the same discontinuity which appears on both results, although at different heights. The discontinuity extended as a layer over south-eastern England from the morning of December 3rd and then rose and drifted away early the following day. On the evening in question a very strong signal, 40 dB above normal, was received in Sussex on a frequency of 180.4 Mc/s from Sutton Coldfield, while communication on 145 Mc/s was achieved between southern England and Germany. From the results of similar analyses carried out daily over nearly a year it is known that the two effects, extended propagation and refractive index discontinuity, are very closely associated.

BOOKS RECEIVED

Department of Scientific and Industrial Research, Annual Report, 1955. Summary of work of all research establishments of the department. Includes notes on the investigation of tropospheric propagation and scattering, direction-finding problems and noise in semiconductor devices by the Radio Research organization. Pp. 321. Price 7s 6d. Her Majesty's Stationery Office, York House, Kingsway, London, W.C.2.

Germanium Diodes, by S. D. Doon. Monograph in the Philips Technical Library popular series on the history, characteristics and applications of crystal diodes. Pp. 85 + viii; Figs. 72. Price 9s 6d. Cleaver Hume Press, Ltd., 31, Wrights Lane, London, W.8.

High Fidelity: The Why and How for Amateurs, by G. A. Briggs. Beginners' guide to the science and art of sound reproduction, covering the equipment required and its handling to produce the best results in the home. The text is enlivened by the author's personal experiences which included the giving of large-scale lecture-demonstrations of high-quality sound on both sides of the Atlantic. Pp. 188; Figs. 65. Price 12s 6d. Wharfedale Wireless Works, Ltd., Idle, Bradford, Yorks.

Hi-Fi Year Book (1956). Edited by Miles Henslow. Survey of current practice in high-quality sound reproduction from discs, magnetic tape and radio. Co-ordinating chapters by acknowledged experts on different elements of equipment are followed by directories of manufacturers. Pp. 136, with numerous illustrations. Price 8s 6d. Miles Henslow Publications, Ltd., 99, Mortimer Street, London, W.1.

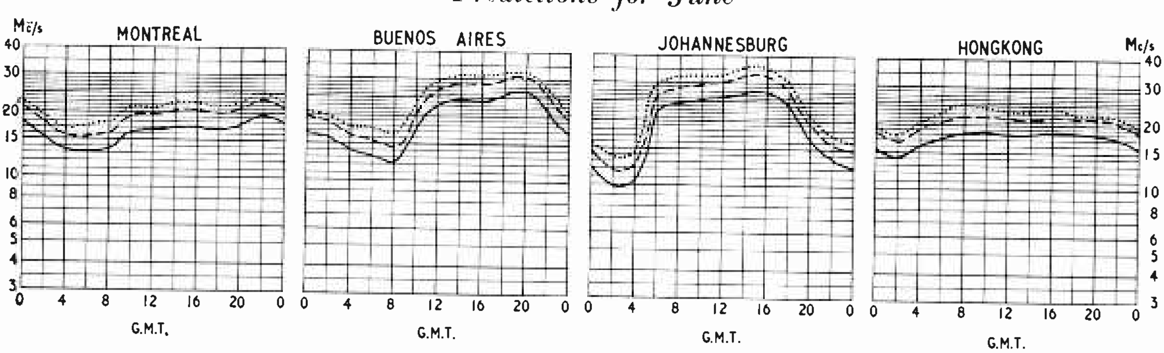
Television and Radar Encyclopædia. Edited by W. MacLanachan. Revised second edition of an illustrated glossary of terms which includes signed contributions from a number of recognized authorities on specialized subjects. Pp. 216; Figs. 224. Price 30s. George Newnes, Ltd., Southampton Street, London, W.C.2.

Radio Servicing Pocket Book, Edited by E. Molloy and J. P. Hawker. Condensed information on test equipment and its use for fault-finding in sound broadcast receivers, including v.h.f. circuits for f.m. reception. Lists the valve sequence and intermediate frequency of popular post-war receivers and includes valve data and valve and battery equivalents. Pp. 200; Figs. and tables 188. Price 10s 6d. George Newnes, Ltd., Southampton Street, London, W.C.2.

Radio Receiver Circuits Handbook, by E. M. Squire. Descriptive analysis, from the practical point of view, of the function of the different stages of sound receivers and amplifiers with a chapter on f.m. discriminators and their associated v.h.f. circuits. Pp. 156; Figs. 122. Price 15s. Sir Isaac Pitman and Sons, Ltd., Parker Street, London, W.C.2.

SHORT-WAVE CONDITIONS

Predictions for June



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 June.

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

·········· FREQUENCY BELOW WHICH COMMUNICATION SHOULD BE POSSIBLE FOR 25% OF THE TOTAL TIME

— — — PREDICTED AVERAGE MAXIMUM USABLE FREQUENCY

———— FREQUENCY BELOW WHICH COMMUNICATION SHOULD BE POSSIBLE ON ALL UNDISTURBED DAYS

Live and Recorded Music

—And Views on Electrostatic Speakers

G. A. BRIGGS' third "adventure in sound" at the Royal Festival Hall last month was described as a concert instead of a lecture-demonstration. His commentary, in consequence, was shorter than on previous occasions, but his audience found it just as fully loaded with wit and wisdom.

This year the London Mozart Players under Harry Blech, Denis Matthews (piano), Leon Goosens (oboe) and Campoli (violin) collaborated in the comparisons of "live" and recorded sound and demonstrated that realism can be achieved in single-channel as well as stereophonic reproduction. Some single-channel sound effects included a recording of a helicopter, which caused many in the audience involuntarily to look upwards.

Moving-coil loudspeakers were used exclusively for this demonstration, and, as if in answer to the unspoken thoughts of some of his listeners, Mr. Briggs had this to say: "In view of the tumult and the shouting created by the new [electrostatic] speakers—or rather by those who make them and listen to them—I cannot let the occasion go by without a brief reference to them. . . . The wide response and freedom from distortion are not in dispute . . . , but we are still waiting for the fanfare to die down and the battle to commence. This is the position as I see it. Electrostatic speakers are coming into use, but the extent to which they will replace moving coils will depend not on perfection in performance but on facility of manufacture and reliability in use under various climatic conditions and overload. Nobody knows the answer yet. Look at pickups. The early models were moving iron and crystal with very crude performance. Then came moving coils and ribbons and frequency-modulation types, giving far superior results; but moving irons and crystals were improved and their position to-day is as strong as ever. In fact, they have already knocked out some of their more fragile opponents. The simplest system always wins in the long run. A similar position applies to microphones."

Responsibility for the amplifier chain and the control of balance and sound levels was once again in the hands (and ears) of P. J. Walker, who also gave a short talk after the interval. Using a one-pound note as a diaphragm, and circulating two coins (one on either side of the paper) to improvise a source of "white" noise of uniform intensity, (it is practically impossible to vary the level by rubbing harder or softer) he showed, with the collaboration of the orchestra, how, by moving the source towards the ear until it could just be heard during a fortissimo passage, a standard of volume level could be carried home in the pocket and used again to set the level of the volume control for realistic orchestral reproduction.

It was intimated that this might be the last of Mr. Briggs' Festival Hall demonstrations. Let us hope that he will be persuaded to give many more, even at the risk of becoming—like the soprano—a "celebrated farewellist."

Pocket "white" noise generator (non-electronic.)

RANDOM RADIATIONS

By "DIALLIST"

"Bournemouth Effect"

A ROMFORD reader sends me an interesting account of a wireless oddity, which he has christened the "Bournemouth Effect!" Some months ago he made a v.h.f. receiver for his parents, then living on the second floor of a seven-storey block of flats in Bournemouth. Excellent reception was obtained with a dipole-*plus*-reflector on the roof. Later, when he was not there, his parents moved to a ground-floor flat in another part of the building. This meant adding 70 feet to the feeder; bringing it up to some 150 feet in all. Low-loss cable wasn't used. When he telephoned to enquire how the set was working under the new conditions, he was surprised by the reply: "Oh, quite well, except that you can hear every train leaving the Central Station." And so, he found, you can indeed! Whenever a train starts to pull out in the daytime the set "huffs and it puffs" like a small locomotive. The effect is not observed at night, when the signal is stronger. My correspondent suggests that each cloud of steam sent aloft by the engines while starting acts for a moment as a screen, reducing signal strength to a level which puts the limiter out of action, and so allowing a "puff" of noise to be heard. He asks whether anyone else has experienced this effect and invites confirmation or refutation of his explanation. I don't think he's far out, myself, but can you think of a better one?

Line-unconsciousness?

AT one time I began to think that the insistence by the man in the street and his wife on bigger and bigger television screens might mean that we should eventually have to abandon our 405-line system in favour of either the French 819 lines or the 625 used by the other European countries. The rooms in the more recently built homes of to-day tend to be on the small side and I found it difficult to see how, in winter time, one could sit near enough to the fire to keep warm and yet far enough from a 21-inch viewing screen to avoid lininess. The answer is that in countless homes you can't. The 17-inch or 21-inch television set is there all right and in chilly weather those who gaze at its screen sit as close as they can to the fire, irrespective of their distance from the set. Careful investigations convince me that the ordinary viewer is becoming—if indeed he has not already become—line-unconscious. He accepts a liny television picture and doesn't notice the lines any more than he notices such little *trifles* as sound-on-vision, violent ringing, or tall people who grow short and thin people who grow fat as they move from one part of the scene to another.

Hi-Fi TV

Many viewers, if not indeed the majority of them, don't worry overmuch about the kind of picture they get, so long as they get a picture. The average eye seems to be just about as accommodating as the average ear and as ready to accept imperfections in reproduction. But more discriminating eyes and ears do notice such shortcomings and are offended by them. I believe that if it were more generally realized how good a 405-line picture *can* be, there'd be as big a boom in high-fidelity television sets as there has been in high-quality sound gear.

E.H.T. Regulation

THE e.h.t. regulation in some television receivers is by no means as good as it might be. In some of them this can produce an effect that may be rather puzzling if you haven't come across it before. The height and width of the image are correctly adjusted on Test Card C, with the black and white borders just fitting into the mask; but when a studio programme starts, you're surprised to find that the picture is too small and has black margins to all four edges. The reason is this. The Test Card contains a good deal of white as well as large areas of pale grey. Reproduction of these makes heavy demands on the e.h.t. supply and when regulation is poor there's a drop (which may be quite considerable) in the e.h.t. voltage. The result of that is an over-large image. If this is fitted into the mask by means of the height and width controls, a normal studio picture, making smaller demands on the e.h.t. supply, won't be enlarged in the same way; the picture will be too small, with black surrounds. Another evil effect of poor e.h.t. regulation is defocusing on whites. If the regulation is really bad this may extend to all the lighter parts of the picture. But don't jump to the conclusion that there is anything more amiss than a faulty e.h.t. rectifier.

The Battle of S.E. London

VERY soon both the B.B.C.'s Crystal Palace station and that of the I.T.A. at Croydon are due to increase their e.r.p. to 120 kW. My heart bleeds not only for viewers in Norwood, Sydenham, Croydon and parts adjacent, but also for harassed servicemen, who will be putting in some pretty work on attenuators and suchlike. As soon as the new transmitter at the Crystal Palace opened up, though on low power, things became more than somewhat hectic. One not uncommon complaint by users of convertors was that when they turned to Band III they got the B.B.C. picture superimposed on that of I.T.A. Just what will happen within short range of a pair of 120-kW stations is, as I write, anybody's guess. Things will, no doubt, sort themselves out fairly quickly; and this experience will give dealers and servicemen some idea of what to expect when the permanent aerial tower at the Crystal Palace comes into use and the e.r.p. of both stations is raised to 200 kilowatts.

Television Interference

IN a recent issue of *W.W.* I wrote that I'd never seen any reference, at any rate in the past few years, to line timebase interference with sound or television reception in any American paper, technical or lay. An Eastbourne reader tells me that he remembers seeing a while ago an advertisement in the *Saturday Evening Post* of a television receiver, which was described as a "Good Neighbor Set" because it was incapable of causing interference with other people's viewing. Well, that's the sole example I've had so far. Another reader reminds me, however, that the real reason for the absence of line timebase interference in the U.S.A. is that the Americans have no long-wave broadcasting band. Complaints of such interference in this country come chiefly from listeners to Droitwich on 200 kc/s; the interference is due to the 20th harmonic of the line timebase which falls at 202.5 kc/s. In the U.S.A. the lowest frequency broadcasting stations are about 550 kc/s and the American line timebase frequency is 15.75 kc/s, so the lowest-order harmonic which can cause interference is about the 35th, which is much weaker than the 20th.

UNBIASED

By FREE GRID

Prudence, not Parsimony

IN the April issue I suggested that those living in remote parts where they cannot get a useful signal from any of the B.B.C. stations might solve their problem by getting the local council to erect a fly-power relay station linked by landline to the B.B.C. network.

A reader writes from Aberdeenshire to say that listeners in his village have overcome their difficulties by private enterprise. As others may be able to solve their listening problems in a somewhat similar way, I quote his words:—

" . . . we are in a blind area so far as the Aberdeen v.h.f. station is concerned and we are, therefore, unable to receive the signals direct. On a nearby hill between us and Aberdeen a signal of some millivolts is obtainable on a simple dipole. We have, therefore, erected an array with aerials on the Aberdeen side which give a gain of about 14 dB. The reflecting area is of rabbit netting, and on our side of the reflector we have aerials which produce usable signals in our village. These aerials are fed from those on the Aberdeen side. . . ."

My correspondent from Aberdeenshire seems to think I have basely impugned the financial habits of the Scots by suggesting in the April issue that co-operation on a basis of voluntary subscription to defray the cost of this sort of thing would not pay north of the border. It would ill become me to foul my own nest, as the great R. L. S. once put it, as I have Scottish blood in my veins, having once received a blood transfusion in a Glasgow hospital.

Prudence, and not parsimony, was what I was trying to imply. My true opinion of the Scottish people was given in the July, 1955, issue.

Historical Heresies

MY QUERY in the April issue, asking when radiotelephony was invented, brought me in a large number of replies. Many of them wrongly placed the invention before 1888 when Hertz first demonstrated the electromagnetic waves that had been mathematically predicted by Clerk Maxwell.

These historical heresies are due solely to the habit of applying the term radiotelephony to systems which employed magnetic or electrostatic induction. Such systems had a very limited range and were never capable of development much further. They were blind-alley systems and therefore analogous in this respect to mechanical methods of television scanning which I condemned in these columns over four years before the B.B.C. followed suit in 1936.

There is one other point which I wish to clear up. Many of my correspondents seek to draw a distinction between the meanings of the words "wireless" and "radio." They argue that "wireless" is an all-embracing word which includes every method of communication without linking wires, including inductive methods, while "radio" implies the use of electromagnetic waves. Needless to say, this is quite wrong. By common usage—an important thing in law—both words mean the same.

Diamond Jubilee

I WONDER how many of you realize that the diamond jubilee of wireless occurs this month. It was on June 2nd, 1896—just 60 years ago—that Marconi filed his first patent. Thirty years later Baird gave his first television demonstration.

If asked in which half of this 60-year period the most progress had been made in radio communication, I think many of the younger readers would say the second because the apparatus of 1926 must seem very crude to them. But to us "old contemptibles" of 1926 the apparatus of 1896 seemed even cruder. Strangely enough, the birth of the famous "Everyman Four" receiver in *Wireless World* in 1926 seemed to mark the turning point between the stone-age and the modern periods of wireless.

I, myself, would unhesitatingly say that the greater progress was made in the first 30 years. After all, in the wider world of general science no single modern invention—not even the harnessing of nuclear energy—has been so great in its effect on the progress of mankind as the invention of the lever and the wheel. I wonder what readers of *Wireless World* 60 years hence will think of us and our much-vaunted technical knowledge? Not much, I expect.

Lèse-Majesté

IN venturing to criticize the words of a writer on which the Editor has set the seal of his approval, I feel much as Nehemiah did when he went into the presence of his lord and master, King Artaxerxes, with "dangerous thoughts" troubling his mind. Nehemiah records that he felt "very sore afraid" but nevertheless by his boldness he won the king over to his viewpoint and maybe I shall do the same.

Let me say at once I am fully in agreement with what V. J. Cooper said in the April issue against the hullabaloo for more lines for both panchromatic and monochromatic television. The one thing I have to criticize is his statement "The question of more lines would probably not be raised if we made the lines invisible by spot wobbling. . . ."

Let me tell Mr. Cooper that spot wobbling is not everybody's cup of tea. According to my observation, the switch controlling the wobbler is, as often as not, set at "off." Spot wobble has the same effect as the soft-focus lens which some professional photographers use to get rid of the lines on the faces of their women sitters. This lens produces a soft fuzzy effect, as it is intended to do.

Suppressing the lines

Information Wanted

RECENTLY I have been busily engaged in searching through learned text books to find the answer to a puzzling little technical problem and have even consulted officials of the Central Electricity Board. I have, however, drawn a complete blank. The problem is this.

If you have been living in a d.c. area and using a universal receiver, and the supply is changed over to a.c., it won't be very long before you need new valves as the heaters soon die. The phenomenon is far more marked in ordinary domestic electric lamps but whether this is due to the fact that they are run at a higher temperature I do not know. If, when the change-over is made, your valves or lamps are fairly new you will have no trouble, but if they have been in use for some time and have therefore become thoroughly saturated with d.c.—if I may so express it—then they will burn out within a few days or even hours.

This phenomenon is not an imaginary one due to faulty observation on my part. It is freely admitted by lighting engineers but nobody seems to know the technical reason for it and that is why I am following the example of St. Paul by appealing to the highest authority which, in this case, is not Cæsar but the learned technologists who read *W.W.* regularly.